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OAO WIDEBAND TRANSMITTER

R. RIPPY

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DAO WIDEBAND TRANSMITTER

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0AO WIDEBAND TRANSMITTER

R. Rippy

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OA0 WIDEBAND TRANSMITTER

INTRODUCTION

This paper is a discussion of a 10 watt 400 MHz FM transmitter developed inhouse by the Radio Frequency Techniques Section of the Space Electronics Branch for OA0-B and follow-on OA0's. The transmitter has a deviation capability of ± 200 KHz, and operates over a temperature range from -20°C to $+60^{\circ}\text{C}$ with a frequency stability of $\pm .0015\%$. It accepts both analog and digital modulation, has a package efficiency of 35 percent and does not require a dc to dc converter.

Basically, the transmitter consists of the functional sections shown in Figure 1.

The modulator generates stable low power 100 MHz sinewave that is frequency modulated by data from the pre-modulation circuits which consist primarily of a bit conditioner and filter. The 100 MHz signal upon leaving the modulator goes

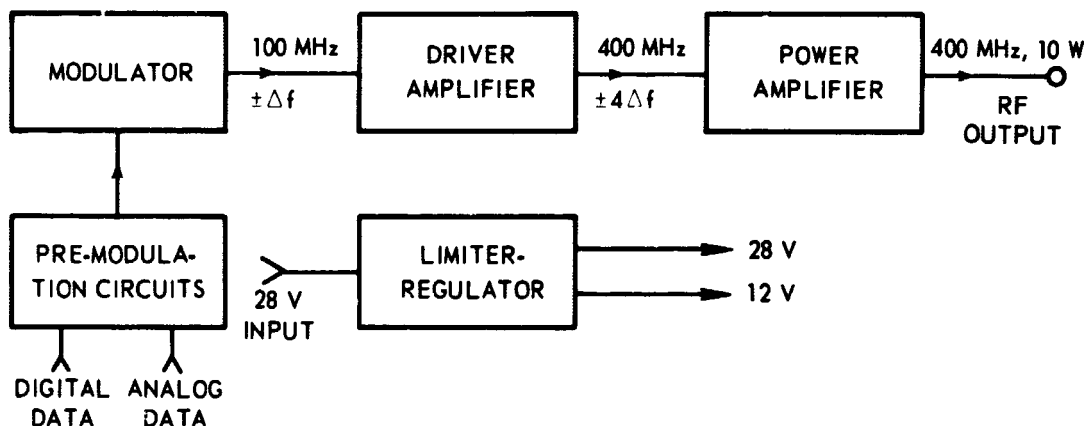


Figure 1. Functional Sections of OA0 Transmitter

to the driver amplifier where it is amplified and the frequency multiplied to 400 MHz. The output of the driver is applied to the power amplifier which produces the required output power.

The limiter-regulator supplies a limited 28 volts to the power amplifier and a precisely regulated 12 volts to all other stages.

These sections will now be discussed in detail in the following order:

(1) Modulator, (2) Pre-Modulation Circuits, (3) Driver Amplifier, (4) Power Amplifier, and (5) Limiter-Regulator. A discussion of RFI/EMC tests is included following the Limiter-Regulator section.

MODULATOR

Purpose

The purpose of the modulator is to generate a stable sinewave whose frequency can be varied linearly by the modulating signal about a nominal 100 MHz. Some of the desirable qualities in a modulator are (1) good short and long term frequency stability, (2) dependability; i.e., immunity from instability caused by temperature changes, electrical and mechanical transients, induced RFI, etc., and (3) linearity with adequate deviation capability and frequency response. This modulator embodies all these desirable qualities.

Circuit Description

A block diagram of the modulator is shown in Figure 2. It consists of a 95 MHz crystal controlled oscillator, a 5 MHz voltage controlled oscillator, a

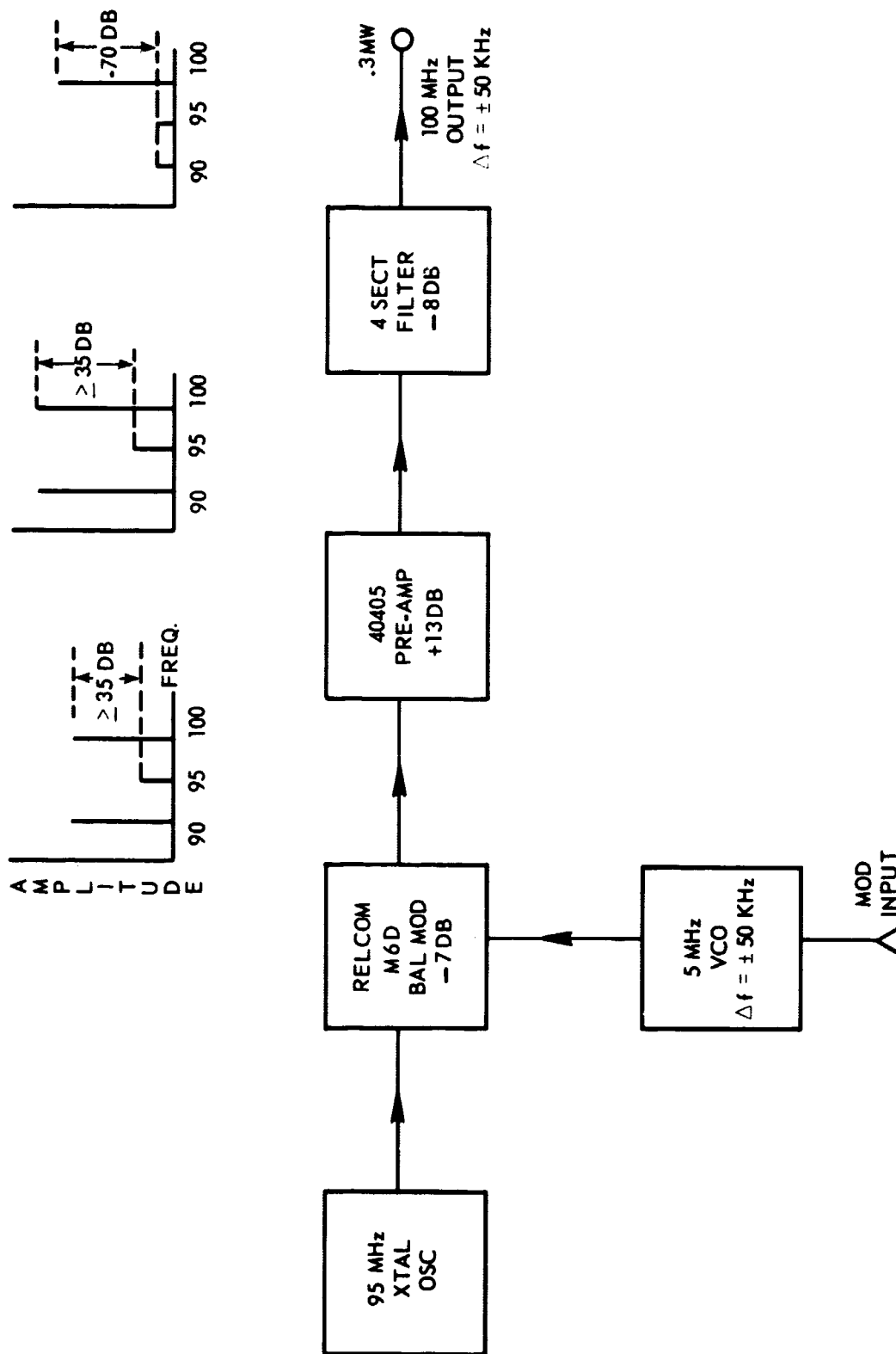


Figure 2. Block Diagram of Modulator Showing Spectrum Output of Each Stage

balanced mixer with a pre-amplifier and a bandpass filter. The outputs from the two oscillators are beat together in the balanced mixer and produce a spectrum at the mixer output that consists primarily of a 100 MHz upper sideband, a 90 MHz lower sideband of equal amplitude and a 95 MHz "carrier" which is attenuated greater than 35 db by the mixer. This essentially double sideband spectrum is amplified and applied to a four section bandpass filter which selects the 100 MHz upper sideband and attenuates all other components. The final output from the modulator is a pure 100 MHz signal having a stability of $\pm .0015\%$ and a deviation capability of at least ± 50 KHz (however, only about ± 16 KHz is required¹). The total harmonic distortion to the modulating wave is less than .5 percent at ± 16 KHz and less than 1 percent at ± 32 KHz.

This modulation technique is but one of many possible methods.² It was chosen because of its simplicity and the absence of phase or frequency locked loops which might lose lock or become unstable during environmental stress. Another technique, "pulling the crystal", is simpler but was rejected because it is inherently a narrowband approach with rather strict limitations on the deviation and modulating frequency response. Also, in order to get the required deviation, a fundamental mode crystal must be used which is less stable than an overtone crystal such as the fifth overtone unit used.

The principal disadvantage of this approach is the relatively bulky filter which must be used to select the upper sideband and the care that must be taken to keep the 95 MHz signal out of the 100 MHz driver amplifier. To solve the

latter problem, it was necessary to mount the crystal oscillator in a completely shielded enclosure (see Figure 12). These two problems become more difficult as the ratio of the crystal to the VCO frequency is made larger. The ratio selected is $\frac{95}{5} = 19$, which is apparently somewhat higher than that usually used in this type of modulator. The reason for using the highest possible ratio, of course, is to obtain a frequency stability approaching that of the crystal oscillator itself.

Performance

The OAO specification requires a stability of $\pm .003$ percent, from -18°C to $+55^{\circ}\text{C}$, however, the design goal was $\pm .0015$ percent from -20°C to $+60^{\circ}\text{C}$. With careful temperature compensation of the VCO, $\pm .001$ percent stability has been achieved. Figure 3 is a plot of frequency versus temperature for flight transmitter No. 002. The maximum deviation from the frequency at ambient is ± 2 , -5 KHz or slightly less than $\pm .001$ percent. The price paid for this stability is primarily the weight and size of the four section filter. Figure 4 is a picture of the filter (a folded tubular helical resonator) and Figure 5 is a typical response curve. The reference or zero db level is 10 db above the top graticule line. It can be seen that the 90 MHz sideband which is 10 MHz below the center frequency will be attenuated about 70 db. The 95 MHz signal will be attenuated 35 db by the filter and 35 db by the balanced mixer for a total of 70 db. The weight of the filter is .36 pound and the insertion loss is 8 db.

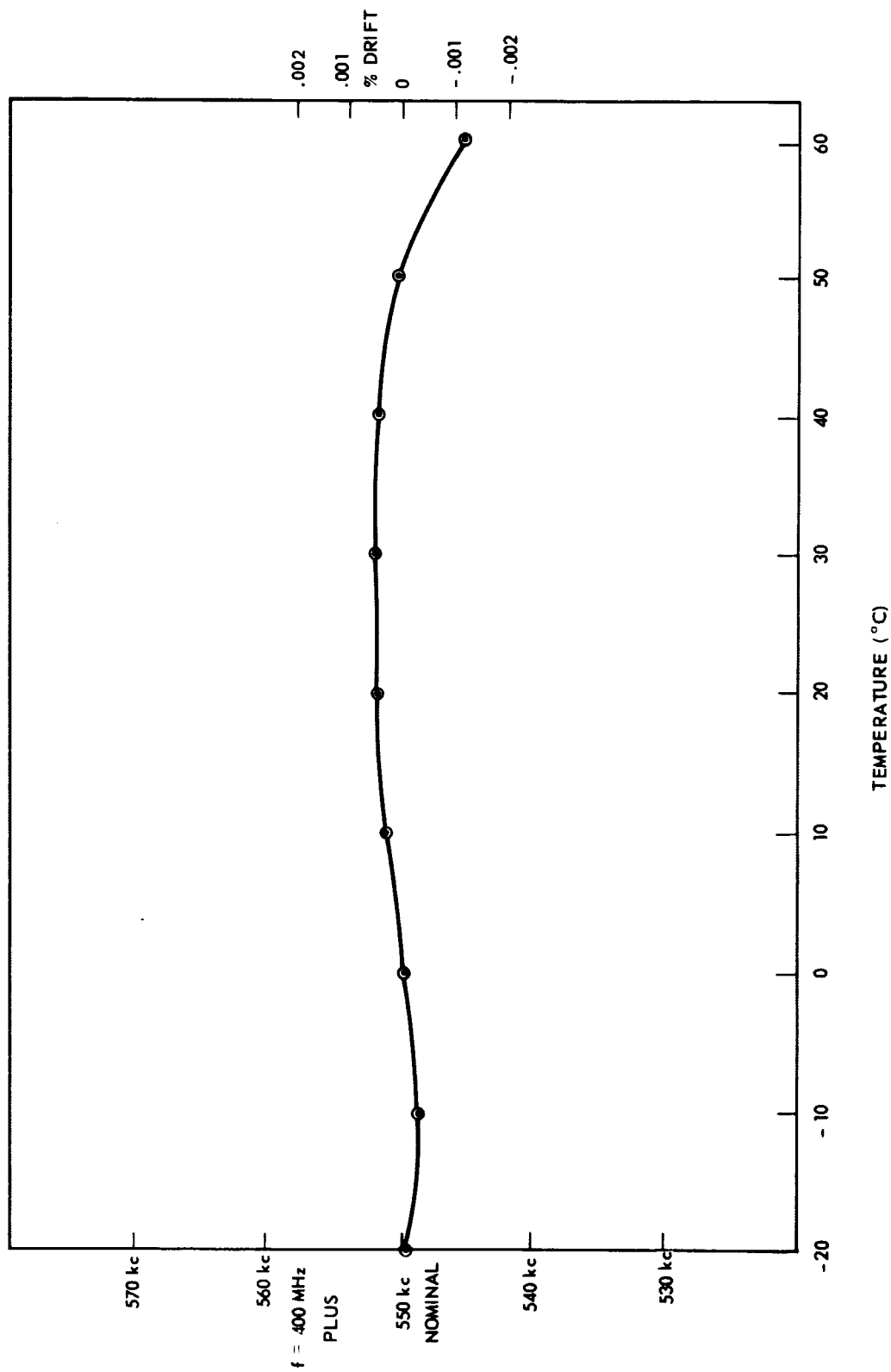


Figure 3. Output Frequency Versus Temperature (°C) For Transmitter No. 002

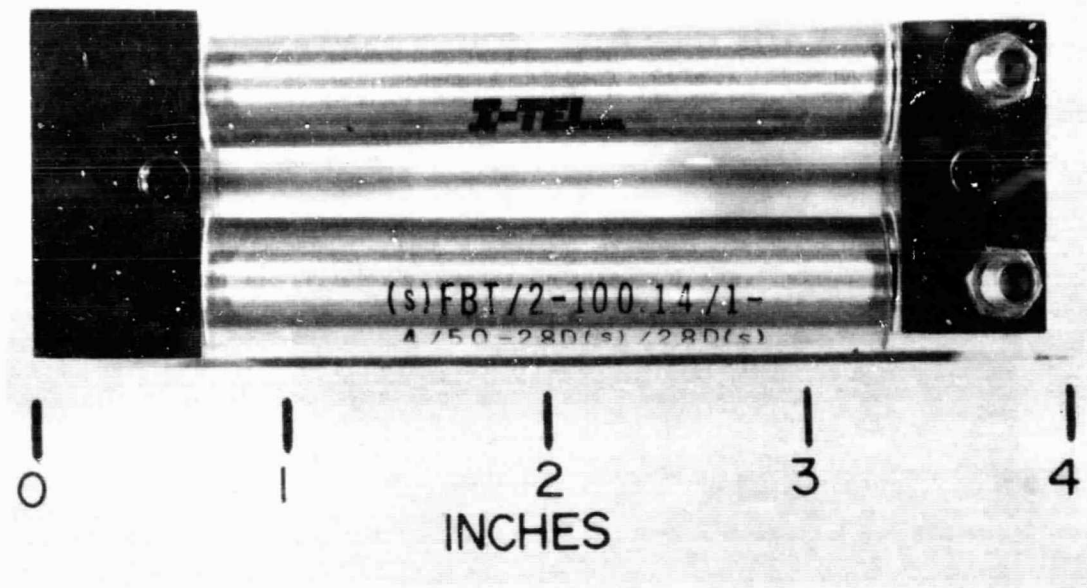
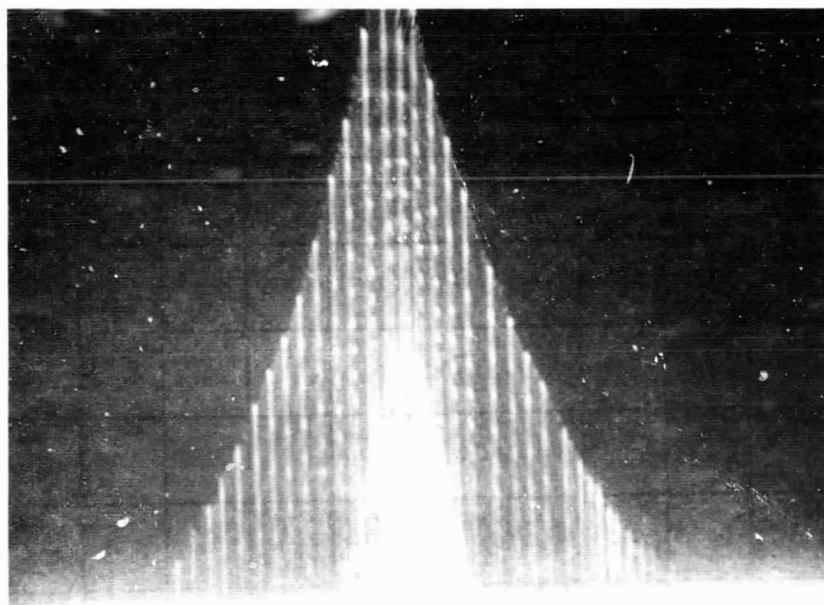


Figure 4. 100 MHz Filter



SCALE —
 VERT.: 10 db/cm
 HOR.: 3 MHz/cm

Figure 5. 100 MHz Filter Response

Unfortunately, the sidebands at the output of the transmitter are not as low as the above analysis predicts because of intermodulation distortion in the driver and power amplifiers. The actual output spectrum at 400 MHz is shown in Figure 6. Here the sidebands are not quite 70 db down. The exact value is somewhat temperature dependent but is usually greater than 60 db.

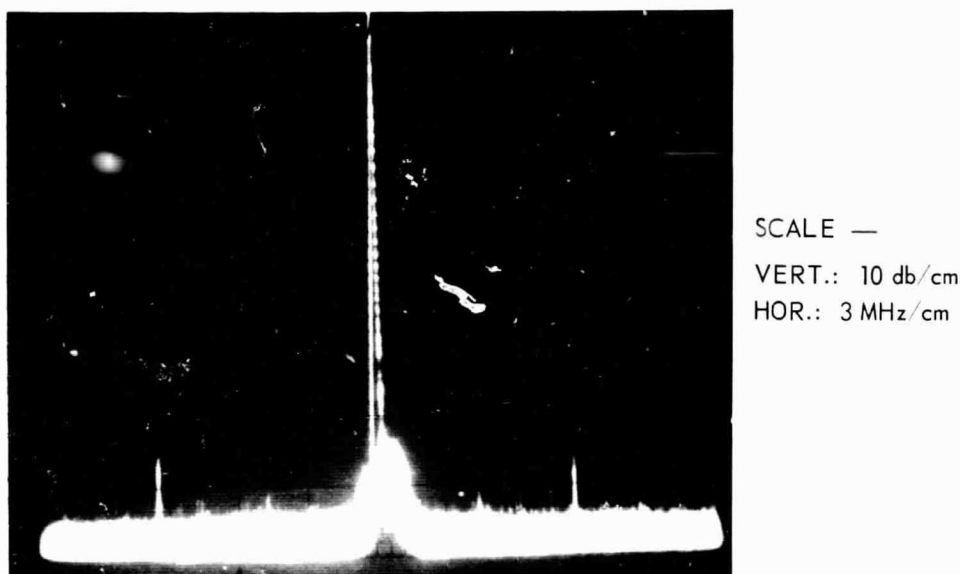


Figure 6. Output Spectrum of Wideband Transmitter Showing Spurious Products of Modulator

Frequency Stability

At this point, frequency stability will be considered in somewhat greater detail. A typical temperature versus frequency characteristic of a 95 MHz fifth overtone crystal is shown in Figure 7. Also shown is a temperature characteristic for a 5 MHz VCO (assumed here to be linear for simplicity) and a curve showing the resultant drift obtained by summing the crystal and VCO curves. It is obvious that if the VCO is made to drift in the opposite direction to the crystal, the drifts

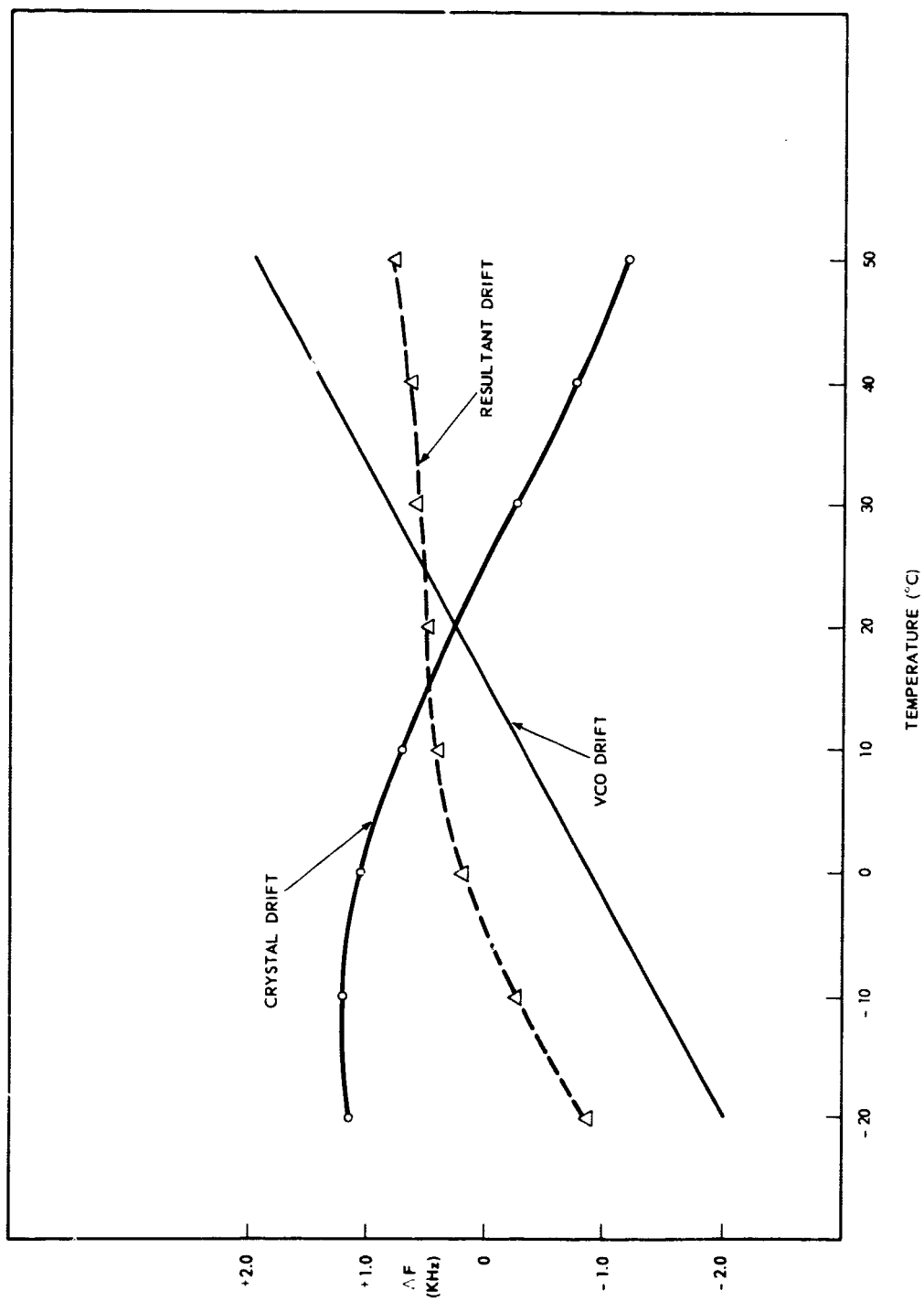


Figure 7. Drift Cancellation 95 MHz Crystal OSC and 5 MHz VCO

will tend to cancel. In the case illustrated, the crystal drifts a total of approximately 2.4 KHz, the VCO 4 KHz, and the resultant roughly 2 KHz at 100 MHz or .002 percent. Thus, if a 5 MHz VCO has a stability of .08 percent and a 95 MHz crystal has a stability of .0025 percent, it is possible for the transmitter to have a stability of .002 percent total or $\pm .001$ percent. Actually the drift in the VCO is more likely to be a rainbow or S shaped curve since temperature compensation tends to remove the linear component. However, these stability figures still apply. Figure 8 shows a set of curves actually measured on transmitter No. 002. Here the stability would have actually been better had the crystal drift been slightly worse because better VCO-crystal drift cancellation would have occurred. The total resultant drift is about 1.2 KHz at 100 MHz or $\pm .0006$ percent. This represents about the best results which have been obtained. A stability of at least $\pm .0015$ percent has been obtained from the breadboard modulator and three complete transmitters. Also, separate tests on the VCO alone corroborate the drift pattern shown. The time required to compensate the VCO to achieve this stability is about one to two days.

Perhaps it should be reiterated that this performance is possible only if one is willing to use a high crystal to VCO frequency ratio and pay the price for filtering which this entails. Theoretically, the VCO drift could be made to exactly cancel that of the crystal. More realistically, however, one can expect to achieve an overall stability approximately equal to that of the crystal alone.

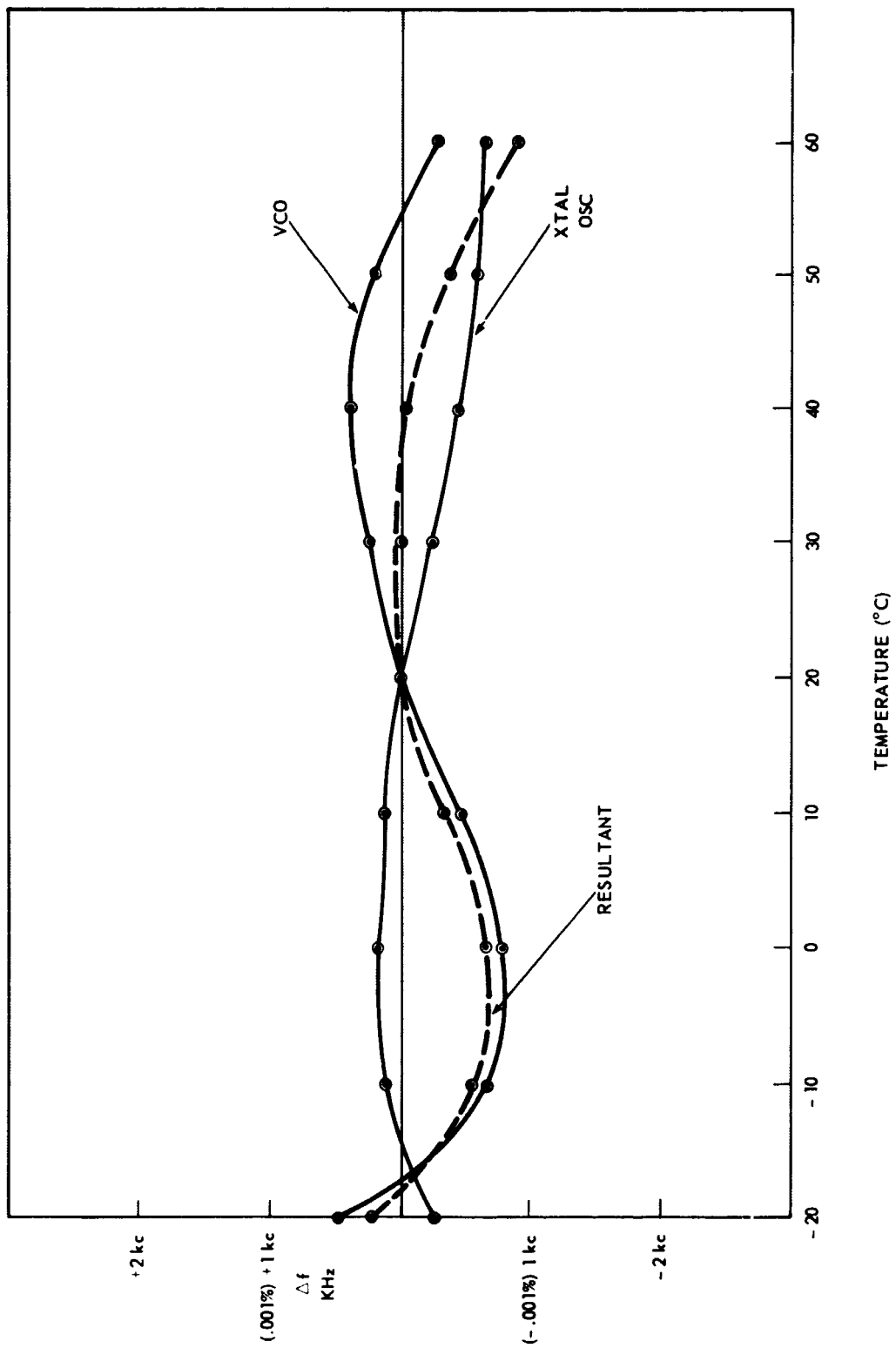


Figure 8. Crystal-VCO Drift Cancellation In Transmitter No. 002

Long Term Stability

The VCO including varactors was operated continuously for a 60 day interval in a clean room environment. The total drift observed during this time was 244 Hz out of a total of 5 MHz or less than .005%. The principal cause of long term drift is evidently

1. Changes in the values of components in the VCO and
2. Varactor bias voltage drift.

Silver mica capacitors having a positive temperature coefficient of 0 to +70 PPM/°C and a drift of $\pm .05$ percent $\pm .1$ Pf over a year are used in the VCO resonant circuit. Erie temperature compensating capacitors having a drift of 1 percent or .5 Pf (whichever is greater) after an accelerated 1000 hour burn-in at 85°C are used to temperature compensate the silver mica capacitors and varactors. A 1 Pf change in capacity in the VCO resonant circuit will cause a 10.8 KHz change in frequency at the transmitter output. In order for the transmitter to meet the design goal of $\pm .0015$ percent frequency stability over a long term period, the multiplied VCO frequency (X4) must not drift more than 12 KHz (assuming no drift or opposite drift from the 95 MHz crystal oscillator and varactor bias voltage). This means the total capacity in the resonant circuit, 920 Pf, must stay within approximately 1.1 Pf or .12 percent of its original value.

The varactor bias voltages are obtained, after filtering, from potentiometer voltage dividers connected across the 12 volt bus. The bus voltage is supplied

by an Amelco 2802BG regulator which is specified to have a typical long term (168 hours) drift of $\pm .05$ percent or 12 millivolts. Since the sensitivity of the VCO is approximately 125 KHz/volt (after X4 multiplication) the frequency drift caused by the regulator may be 1.4 KHz each for the analog modulation and frequency adjust varactors or a total of 2.8 KHz. There is a third varactor for digital modulation but it is reversed biased in the unmodulated state into the insensitive region of its V versus C curve and is not likely to contribute significantly to long term drift.

Obviously, from the above discussion, extremely high quality components and quality fabrication must be used in the VCO if long term drift within the $\pm .0015$ percent range is to be maintained. The sensitivity of frequency to capacity is so great that ordinary printed circuit techniques could not be used on the VCO board. The frequency determining components had to be mounted above the board on standoffs and point to point wiring used to connect them. With the components lying on the board and connected in the conventional manner by metalization on the bottom of the board, the frequency would not stabilize at low temperatures or retrace the same path on successive temperature cycles. Apparently, moisture absorbed by the board caused a sufficient change in stray capacity to produce this behavior. At any rate, the problem disappeared after mounting the critical components above the board and employing point to point wiring. Figure 9 shows a close-up of the VCO board illustrating the component mounting arrangement.

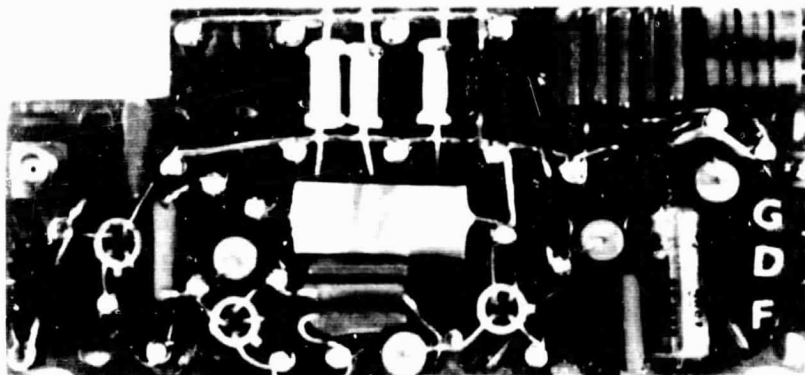


Figure 9. VCO Board Showing Critical Components Mounted on Standoff Insulators

The problem of obtaining long term stability in the ± 0.0015 percent range over large temperature and time excursions is very difficult, and although stability of this magnitude has been achieved repeatedly over an 80°C temperature range, further tests on the complete transmitter are needed to determine if the specification can be met over a period of months or longer.

Short Term Stability

The preceding discussion has dealt with slow variations in frequency caused by component aging and changes in temperature. Variations that occur in a time interval of one second or less, i.e., short term stability,^{3,4} will now be considered. These variations are primarily due to noise generated by the resistive component of the crystal and VCO resonant circuits and the transistors in these circuits. Where frequency control varactors are involved, an important contribution is made by noise on the bias voltage and induced noise from nearby circuits. The best possible short term stability would exist if the transmitter were

100 percent crystal controlled and used as varactors. Since it is only 95 percent crystal controlled and contains varactors, some degradation is to be expected.

The photographs in Figure 10 illustrate the short term stability of transmitter No. 002 in comparison to a purely crystal controlled 400 MHz signal. In each photograph the transmitter spectrum is on the left and the crystal spectrum on the right. Photographs A-B, C-D, and E-F represent the transmitter and the pure crystal spectrum multiplied to 2.4, 4.0, and 10.4 GHz respectively. Two

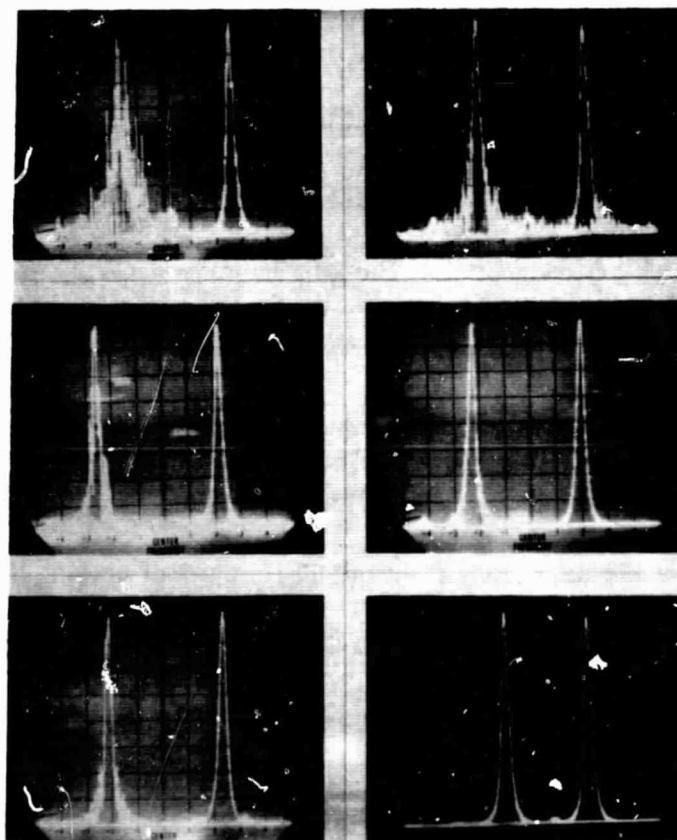


Figure 10. Short Term Stability of Wideband Transmitter Compared to a Purely Crystal Controlled Signal at 2.0 (Bottom), 4.0 (Center) and 10.4 (Top) GHz.

Left Photos: Normal Varactor Bias. Right Photos: Varactors Highly Reversed Biased.

pictures at each frequency are shown to illustrate the adverse effect varactor bias voltage noise has on the spectral purity. The left photograph in each figure was taken with normal operating bias voltage on each varactor, and the right photograph with the varactors highly reversed biased to decrease their sensitivity. It is obvious that a significant part of the instability is caused by noise voltage on the varactors.

An approximate calculation of the noise voltage can be made from the pictures. First, we must find the frequency deviation Δf caused by the noise band of interest. Then, we may multiply this Δf times the known transmitter modulation sensitivity in V/Hz to obtain the noise modulating voltage. There is a noise spike in the 10.4 GHz photograph (Figure 10E) which is about 200 Hz and 12 db below the "carrier" (i.e., center of the spectrum). Since the spectrum analyzer IF bandwidth is 50 Hz, this component is really due to the total noise in a 50 Hz band centered 200 Hz below the center of the spectrum.

Let us calculate the equivalent voltage of a 200 Hz sinewave that will produce a spectral line of the same height as the 50 Hz band of noise. If the modulation index is much less than 1, the sideband-to-carrier voltage ratio is equal to one half the modulation index: $\frac{V_s}{V_c} = \frac{1}{2} \frac{\Delta f}{f_m}$. Substituting $\frac{V_s}{V_c} = .25$ (12 db) and $f_m = 200$ Hz, we get $\Delta f = \pm 100$ Hz at 10.4 GHz. Because the modulation index here is one half, and not much less than 1, this calculation is only approximate, but should give the correct order of magnitude of the noise voltage. A Δf of ± 100 Hz at 10.4 GHz is equivalent to a Δf of ± 3.84 Hz at 400 MHz. The sensitivity of the

modulator is approximately $\pm 62.5 \frac{\text{KHz}}{\text{Volt}}$. * Therefore, the sinewave voltage which would produce a spectral line the same height as the one in question is

$$\frac{1 \text{ V}}{6.25 \times 10^4 \text{ Hz}} \times 3.84 \text{ Hz} = 61.4 \text{ microvolts.}$$

The same voltage at a lower modulating frequency would produce a larger sideband because the modulation index would be greater.

In order to minimize the noise voltage conducted into the transmitter while the stability measurements were made, all modulation leads were cut short and grounded to the chassis. Apparently, much of the noise which appears in the photographs originates within the transmitter. The noise on one Amelco 12 volt regulator was measured at 100 microvolts with a 10 microfarad capacitor connected across the output. Wideband noise on a zener was measured at 1 millivolt. This points out the importance of biasing the varactor with a very stable, "quiet" voltage, shielding modulation leads from nearby AC leads, and using the least sensitive modulator practical. Had the varactor used, for example, had a sensitivity of 10 volts per 62.5 KHz, instead of 1.0 the effect of the noise would be reduced by a factor of 10. The transmitter design reflects a tradeoff between short term stability and modulation linearity. A large modulating voltage is desired for good short term stability, but a low modulating voltage is desired to make good linearity easier to obtain. Actually in an FM transmitter, it is somewhat

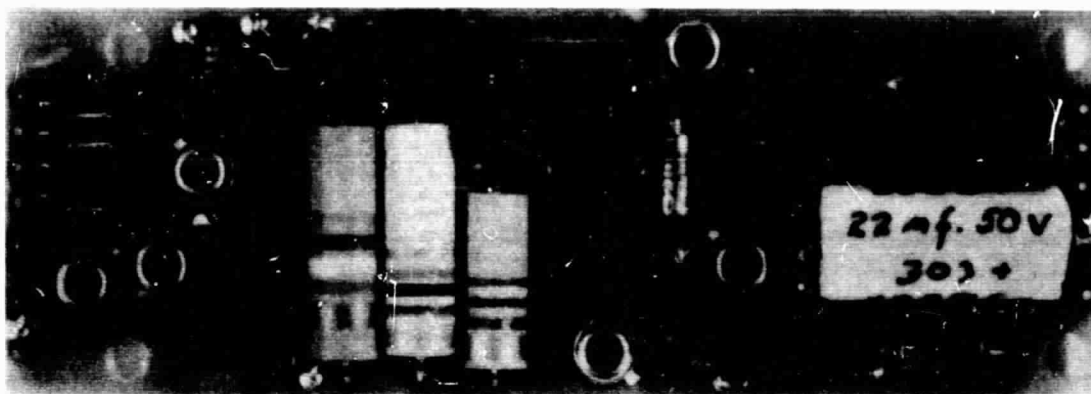
*This figure applies at the varactor only. The sensitivity at the analog modulation input terminal is $\pm 62.5 \text{ KHz}$ for a 4 volt PP input.

academic to strive for the ultimate short term stability. A more realistic criterion for determining the adequacy of short term stability is perhaps that the noise due to spectral impurity be one tenth that due to distortion introduced by the modulator. For example the OAO Specification allows 3% analog distortion referred to ± 62.5 KHz, therefore, it should be satisfactory for the transmitter to have a total residual noise of .3% or 188 Hz. This is what has been measured. However, it represents the residual noise in the test equipment rather than the transmitter. The residual noise in the transmitter is believed to be considerably less than .3%.

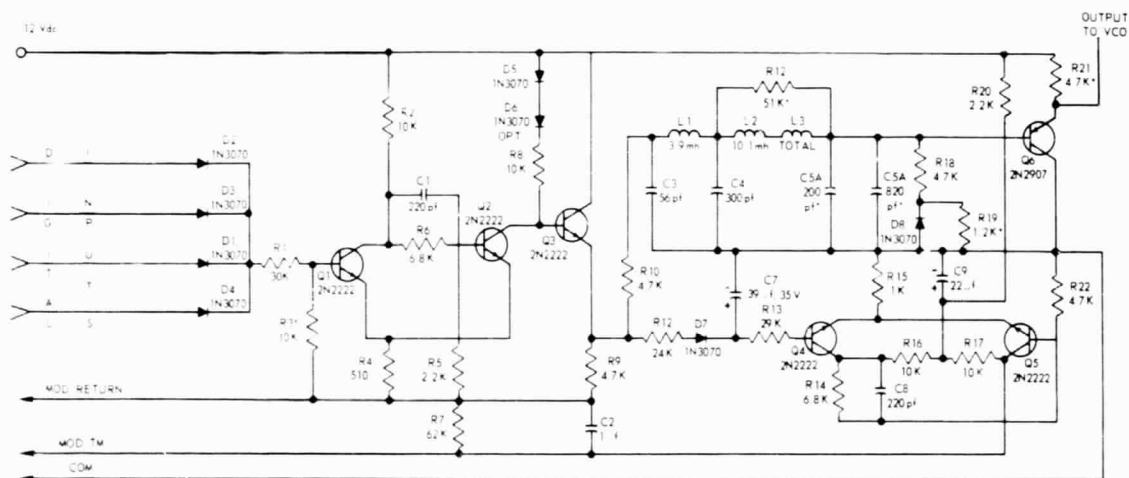
PRE MODULATION CIRCUITS

Digital Filter

The purpose of the pre modulation circuits is to process the analog and digital input data for application to the varactors. The digital processing circuit consists primarily of a schmitt trigger which reshapes the input waveform and a filter which attenuates the harmonics of the higher frequencies so that the transmitted spectrum is properly contained within the allotted bandwidth. All digital processing circuits are mounted on a single printed circuit board which is illustrated in the photograph and schematic of Figure 11. There are four digital inputs each of which is applied to the schmitt trigger (Q_1 and Q_2). Data is always applied sequentially to one input at a time and the inputs are isolated from each other by diodes D_1 through D_4 . A digital one is represented by a 5 to 18 volt



A



B

Figure 11. Digital Input Filter

input and produces a deviation of -100 KHz. A digital zero is represented by a 0 to 4.0 volt input and produces no deviation. The schmitt trigger prevents any change in deviation due to amplitude instability in the modulating waveform or difference in level between the four inputs. The reshaped wave at the output of the schmitt trigger feeds emitter follower Q_3 . The premodulation filter is connected between the output of Q_3 and the input of a second complimentary emitter follower Q_6 , which isolates the filter, partially cancels dc drift, and provides a low output impedance to drive the varactor on the VCO board. Diodes D_5 , D_6 , and D_8 are used for dc drift cancellation.

Digital Modulation

The digital modulation process is not linear, i.e., the digital deviation is not a linear function of modulating voltage. Consequentially, if the voltage at the output of the low pass modulation filter drops 3 db, the deviation, due to varactor non-linearity, may drop the equivalent of 4 db or more. This complicates filter design and has made it necessary to empirically determine the 3 db bandwidth of the filter alone that will give the desired modulation bandwidth. As an example, a filter with a 3 db bandwidth of 90 KHz was found to have a true transmitted 3 db bandwidth of 68 KHz. The true bandwidth is measured at the output of a receiver or on a spectrum analyzer. The filter skirt response is also distorted although not as severely since when the the filter output voltage has dropped appreciably, the varactor is operating in a more linear region of its characteristic curve.

The above problem could have been avoided had the digital modulation process been made linear by using a varactor of the same type used to obtain analog modulation. However, this varactor requires less than one volt to produce a 100 KHz change in frequency at 400 MHz. The voltage now required to obtain this deviation is 5.5 volts which is large enough compared to the dc drift in the schmitt trigger and emitter followers that the frequency drift caused by these circuits is insignificant. This benefit would likely not occur had the linear circuit been used. The drift problem does not exist in the analog circuit because it is ac coupled.

One auxiliary circuit, also on the digital input filter board, is a second schmitt trigger (Q_4 and Q_5). When data is applied to any digital input, the circuit triggers, causing a voltage rise at the collector of Q_5 which is used to signal the ground station that digital data is present.

Analog Modulation

The transmitter has one analog pre-modulation circuit which is located on the VCO board. It is extremely simple in that it contains only a potentiometer to control the modulation sensitivity. The raw analog data is applied directly to the varactor through an RF choke without any processing. This is possible because of the rather unusual characteristic of the analog varactor, an MV1403. The capacity varies inversely with the square instead of the square root, as is usually the case, of the applied voltage. If this varactor is connected in series

with the correct fixed capacitance, the series combination capacitance will vary linearly with the varactor voltage over a limited operating region (see Appendix). The VCO frequency, although a function of $\frac{1}{\sqrt{C}}$ nevertheless varies quite linearly with C, if C only changes a small amount. The result is that the VCO frequency will vary linearly with varactor voltage over about a ± 50 KHz frequency range, which is more than needed. The frequency response is about 2 Hz to 67 KHz at the 3 db points and may be extended from DC to well over 100 KHz if desired.

DRIVER AMPLIFIER

Power Levels

The driver receives a .3 milliwatt 100 MHz FM signal from the modulator and amplifies it to a power level of 140 milliwatts. This signal then drives a X4 step recovery diode multiplier which has an output of 90 milliwatts at 400 MHz. The spectrum from the multiplier contains harmonics of the 100 MHz input which may be only 15 db below the desired signal. In order to eliminate these spurious frequencies, a 400 MHz bandpass filter is connected between the driver output and the power amplifier input. After passing through the filter, the power level is reduced to approximately 50 milliwatts which drives the first stage of the power amplifier. All harmonics at the output of the filter are at least 70 db down. A picture and schematic of the driver are shown in Figure 12. Also shown in the picture are the crystal oscillator enclosure and the transistor used for overvoltage protection to be discussed in a later Section.

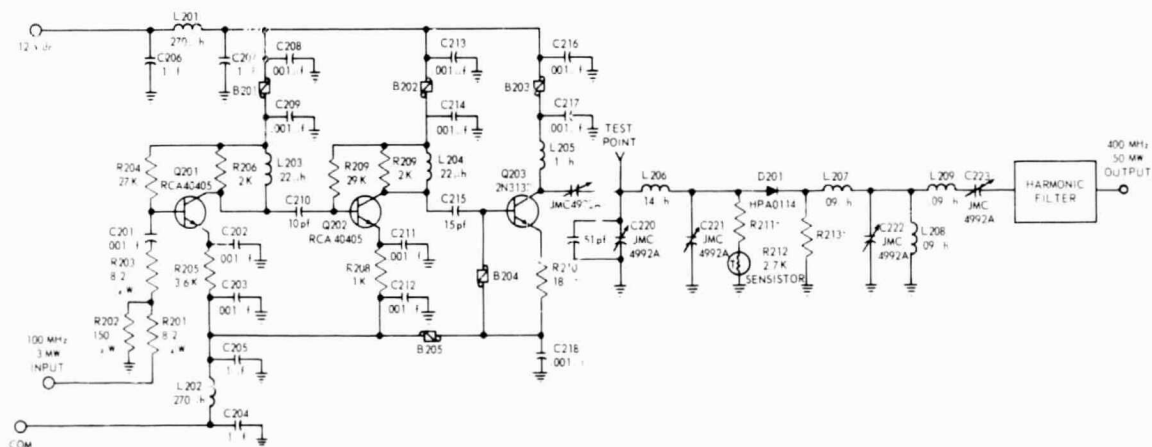
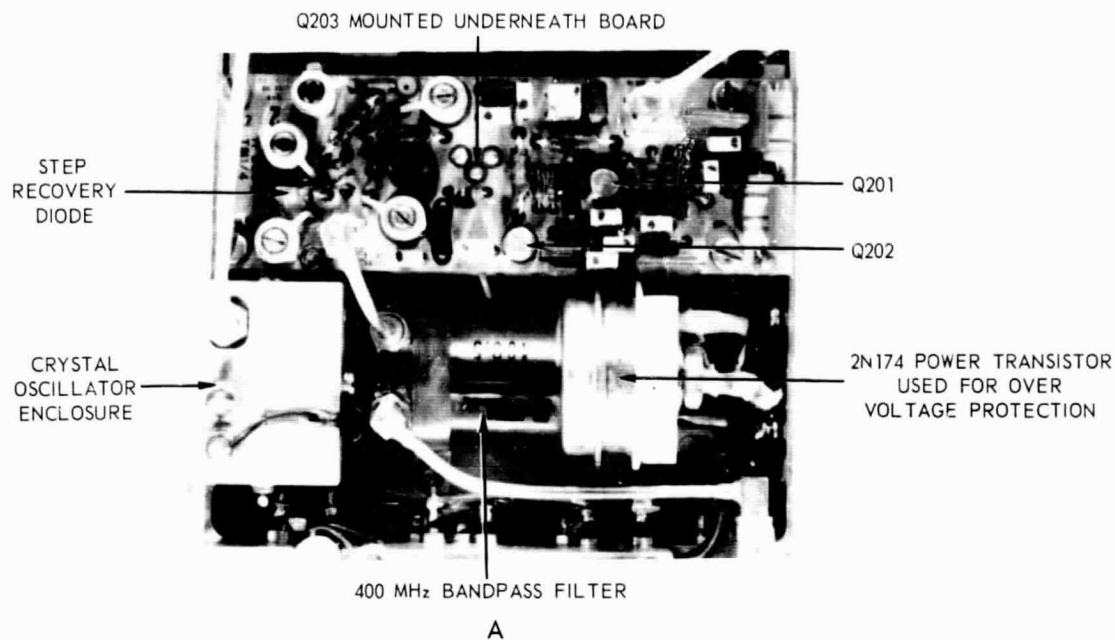


Figure 12. Driver Amplifier and Harmonic Filter. Picture Also Shows Crystal Oscillator Enclosure and Power Transistor Used For Overvoltage Protection

The gain of each transistor stage is approximately 10 db. Thus, a .3 milliwatt signal at the input passes through the 3 db pad and arrives at Q_{201} as .15 milliwatts. The output of Q_{201} is 1.5 milliwatts, of Q_{202} , 15 milliwatts, and of Q_{203} , 140 milliwatts. No input matching is employed to the 40405's. If biased correctly the input impedance is so near 50 ohms that none is required. The total current drain is 35 milliamps at 12 volts.

Components

As shown, the entire circuit excluding the filter is built on a printed circuit board. A problem was encountered in heatsinking the 2N3137 which must dissipate about 80 milliwatts. It was decided to provide at least some means other than radiation for eliminating the heat. The method chosen was to mount the transistor upside down underneath the circuit board so that when the board is installed, the transistor fits into a finger-grip heatsink which is attached to the chassis.

The components in the schematic which are labeled with the prefix B, are ferrite beads used to decouple RF from the power leads. They consist of three turns of wire on a small torroidal core and have a purely resistive impedance of about 600 ohms at 100 MHz. B_{204} is used as a low Q dc return for the base of the 2N3137 (Q_{203}).

Q_{201} and Q_{202} were formerly 2N918s but were changed to 40405s because the latter has a more nearly constant power output over the operating temperature

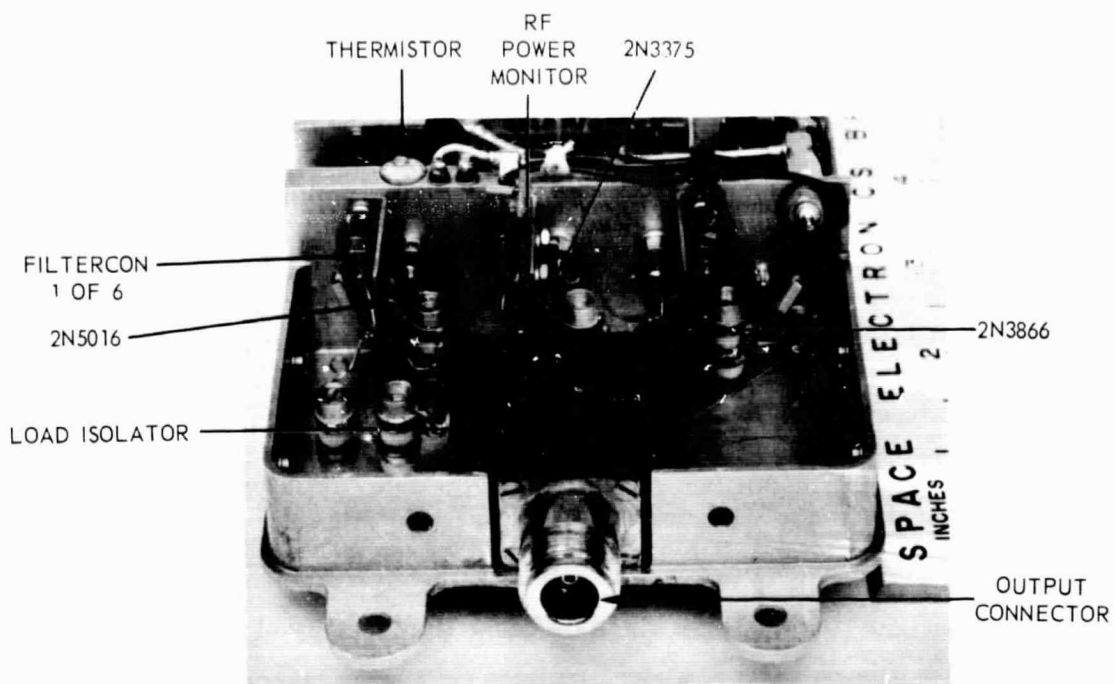
range. The parameters of both devices are sufficiently similar that no circuit modifications were required. It is important to maintain the drive level to the X4 multiplier and power amplifier as nearly constant as possible in order to prevent parametric oscillations (to be discussed later). The wider the temperature range over which the transmitter must work, the more critical the problem becomes. Prior to replacing the 2N918s, it appeared virtually impossible to make the qualification transmitter work over the -35°C to $+70^{\circ}\text{C}$ qualification temperature range without some jump or breakup. R_{206} , R_{209} , and the 3 db pad at the input of Q_{201} are components which increase the stability of the amplifier by resistive loading.

POWER AMPLIFIER

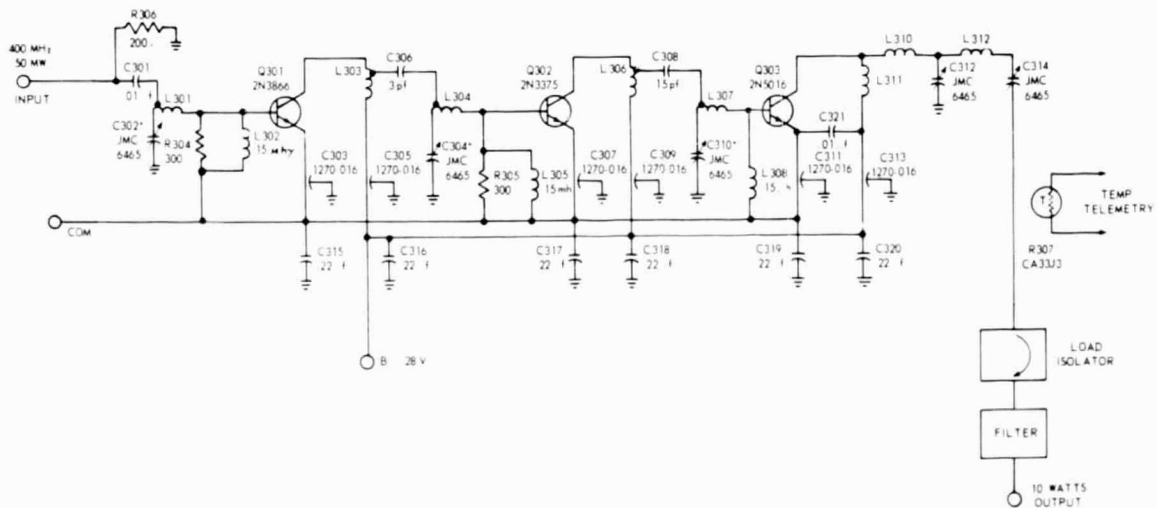
General

The RF power is generated in a three stage Class C amplifier which produces 12 watts output with 50 milliwatts input at 400 MHz. The output transistor is protected from load mismatch by an isolator which has a .7 db insertion loss, reducing the power at the output connector to 10 watts. This can be raised to as much as 13 watts, but the amplifier is set up to deliver 10 watts in order to provide a safety margin for the transistors. All stages operate at 28 volts. A schematic and picture of the amplifier is shown in Figure 13.

The schematic illustrates a rather unusual feature of the power amplifier; both power leads are isolated from case ground. This is necessary to satisfy a



A



B

Figure 13. Power Amplifier

spacecraft requirement that all power and telemetry leads be isolated from the case. The usual method of meeting this specification is to use a DC to DC power converter. However since isolation is achieved inside the transmitter and since a regulated 28 V bus is available in the OAO, a power converter is not needed. This results in a power saving of 7 watts and an improvement in reliability. Protection from occasional positive and negative spikes or even sustained over-voltage on the regulator bus is provided however. This is discussed later in the Section on the regulator.

Note in the picture there are two filtercons at each stage for isolation of both power leads. Looking from left to right, the transistors are 2N5016, 2N3375, and 2N3866. The emitter of the 2N5016 is bonded to the TO-60 case inside the header which makes it necessary to insulate the case from ground. Although not visible in the picture, this is done with two Beryllium Oxide washers, one underneath the case and the other underneath the nut behind the heat sink partition. Lifting the transistor off ground resulted in a severe stability problem until a .01 MFD capacitor was soldered between the transistor case and the body of the collector filtercon. Apparently, the RF ground provided by the filtercon connected to the emitter pin was not adequate.

A thermister is imbedded in the heat sink partition just above the 2N5016 to provide telemetry data on the chassis temperature. Telemetry data on output power is also provided by rectifying part of the RF voltage at the load isolator.

The rectifying circuit is on the small printed circuit board mounted on the center shield of the power amplifier.

A breakdown of the individual stage currents and power levels for a typical transmitter at room ambient is given below.

Stage	Current (MA)	DC Power In (w)	RF Power Out (w)	DC to RF
				Efficiency (%)
2N3866	55	1.5	1	65
2N3375	260	7.3	4	55
2N5016	630	17.6	12	68
All PA	945	26.4	12	45
All Other Stages	75	2.1	.05 (Driver Output)	2.37
TOTAL	1020	28.5	12	42

A 2 watt loss in the load isolator reduces the total package efficiency to $10/28.6 = 35\%$.

Depending upon how each individual transmitter is tuned, the efficiency may be as high as 38 percent or as low as 31 percent. This results from having to make a tradeoff during the tuning process between maximum efficiency and maximum stability. The efficiency of a well tuned transmitter varies between 34 and 38% as the temperature varies between -20°C and 60°C.

Without the output load isolator, the package efficiency would be greater than 40 percent. However, it is believed that the protection against parametric instability and damage to the output transistors provided by this device justifies the cost in efficiency. It provides greater than 20 db attenuation to reflected power and will safely dissipate up to 20 watts into an internal termination. As a result of the isolator, the power amplifier will not be damaged or break into parametric oscillation when connected to a load of infinite VSWR at any phase angle. The main disadvantage of the device is the insertion loss, which at .7 db is rather high. However, isolators have recently been purchased which have an insertion loss of only .4 db. If these units were used, the power output would increase to 11 watts and the overall efficiency would be about 38 percent.

Harmonic Output

The amplifier second harmonic content at the output of the isolator is about -50 db down from the carrier level. To obtain greater attenuation, a short tubular low pass filter that mates to the output connector on the load isolator may be used. With the filter installed, all harmonics are greater than 75 db down from the carrier level. Incorporating the filter inside the transmitter is a future design goal.

Parametric Instability

A major design problem is elimination of parametric oscillation in the transmitter. This type of instability often referred to as spectrum breakup, squegging, relaxation oscillation, motorboating, etc., is the leading problem encountered with the power amplifier.

It affects virtually all large signal RF transistors and varactors. Unfortunately, the behavior is perfectly normal for these devices. It has been shown^{5,6} that a resonant circuit containing a voltage variable reactor such as a varactor or transistor junction has unstable regions where the derivative of amplitude of the resonance curve with respect to frequency is infinite. At such points, the amplitude will jump abruptly to a new point on the curve, and may then jump back again. Such behavior is parametric oscillation.

Figure 15 shows a picture of a transmitter output spectrum which has broken up due to parametric oscillation. The sidebands are caused by the oscillation (although they could easily be mistaken for legitimate modulation) which produces a combination of AM and PM at a frequency which usually varies as the transmitter

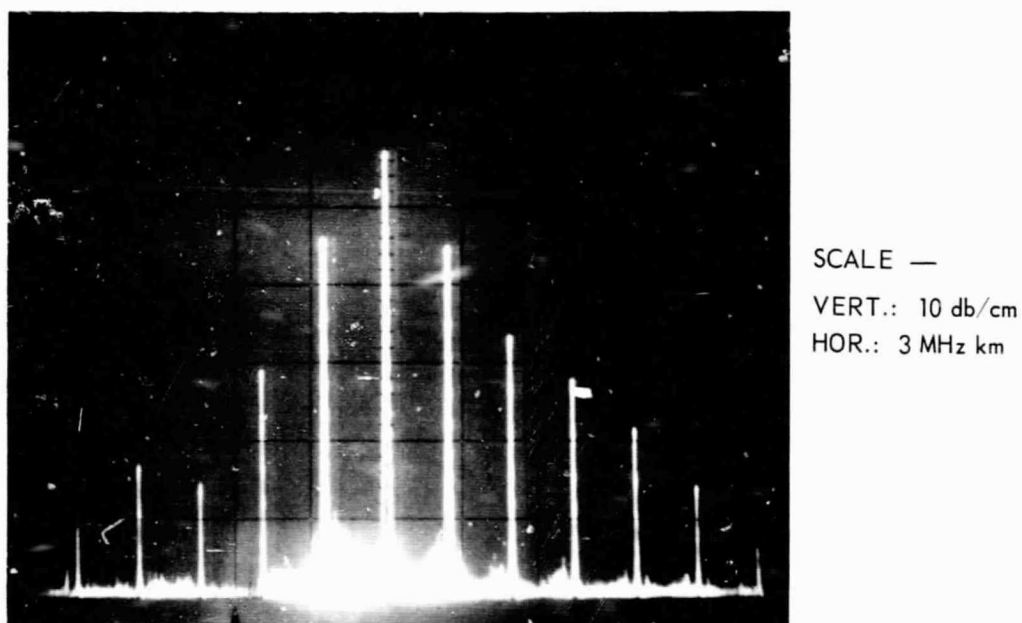


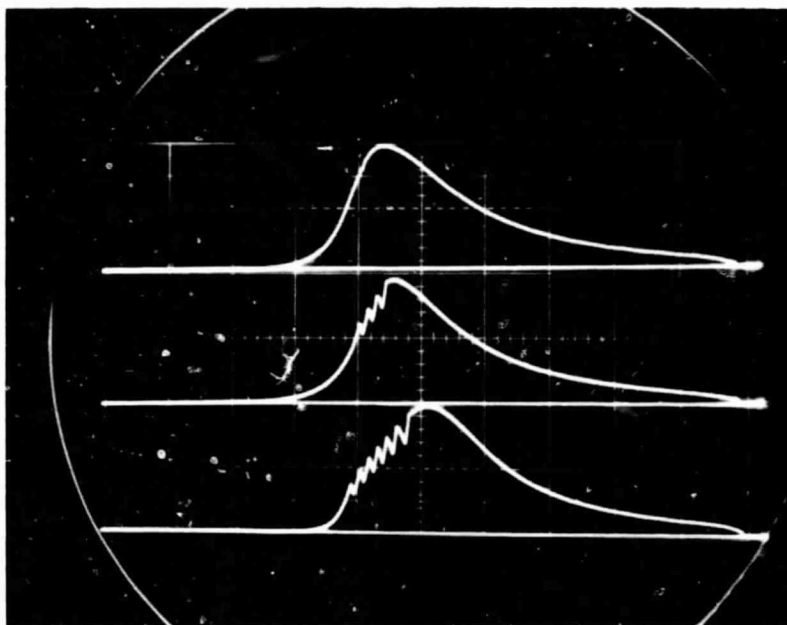
Figure 15. Parametric Oscillation in Frequency Domain

is tuned. The AM component can often be observed on the power leads of the oscillating stage.

This same phenomenon at times appears as an abrupt jump in output power. Either the jump or oscillation may occur as a result of subjecting the transmitter to changes in temperature, VSWR or internal drive level. The disturbance may also appear or disappear as a result of moving a component or bending a lead in the power amplifier or driver. When a change in temperature is the precipitating factor, the instability usually disappears when the temperature is returned to its original value. Anytime the output power changes abruptly or the spectrum appears distorted only to later clear itself and be normal, parametric oscillation should be suspected.

One of the most effective means of suppressing the oscillation is to resistively load the resonant circuits which are associated with the varactors or transistors involved. This may be done by actually adding a resistor or by designing the matching networks with the least possible Q to achieve the required impedance match. The latter method must usually be used in high power stages unless special networks are used which resistively load the circuits only at out-of-band frequencies.

Figure 16 illustrates the effectiveness of resistive loading in quenching the oscillation. The effect in the photograph was produced by applying a swept frequency to a simple varactor tuned parallel resonant circuit. The horizontal axis



SCALE —
VERT.: AMPLITUDE
HOR.: FREQUENCY

Figure 16. Parametric Oscillation Illustrating How Resistive Loading (Upper Traces) Quenches the Oscillation

is frequency and the vertical axis is amplitude of the voltage appearing across the circuit. The bottom trace was made with no circuit loading and the upper ones with successively greater loading. Figure 17 illustrates the jump phenomenon and how heavier loading (upper traces) reduces the severity. Both forms of the instability are difficult to completely suppress if a transmitter must work over a wide temperature range. The following measures have been found helpful:

- (1) minimum Q matching, (2) resistive loading, (3) output load isolator, (4) inter-stage isolators, (5) low frequency by-passing of dc leads, and (6) careful tuning.

The power amplifier incorporates all of these methods except number 4.

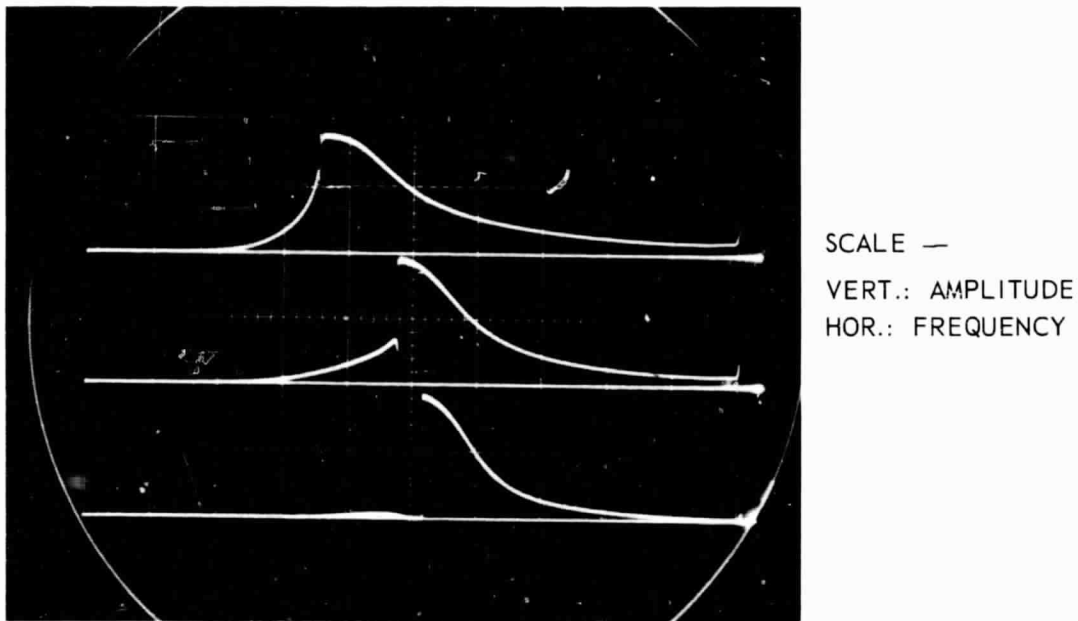


Figure 17. Jump Phenomena, Upper Traces Show Effect of Resistive Loading

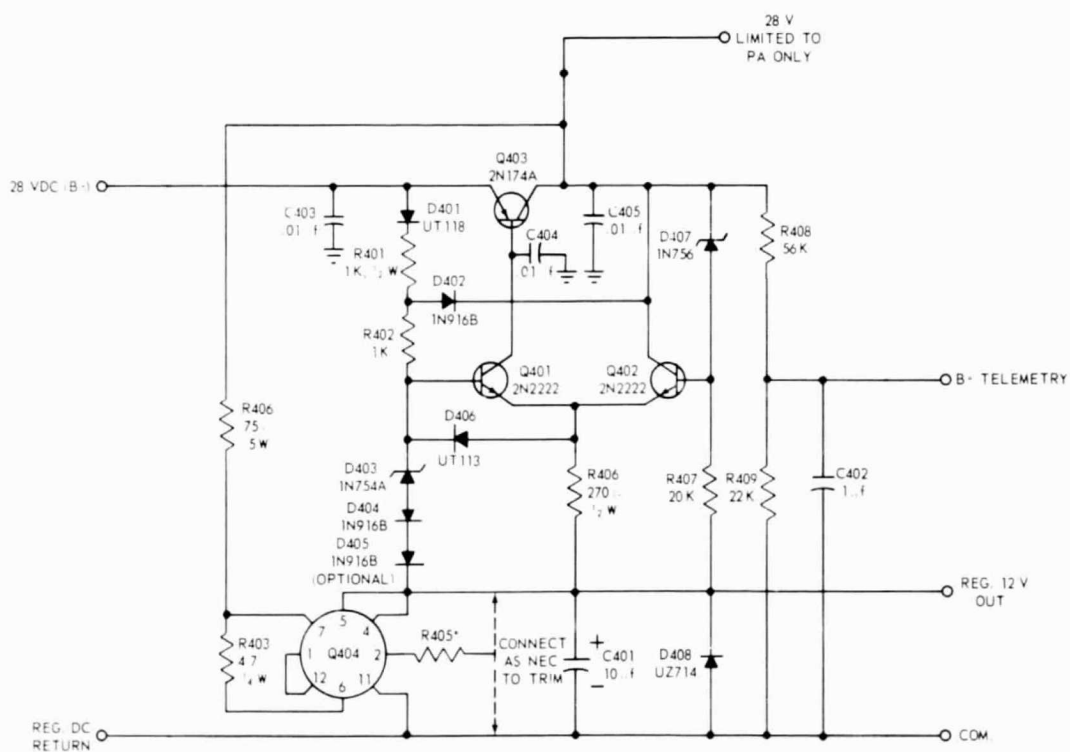
REGULATOR-LIMITER

Over Voltage Protection

The transmitter operating voltage is a nominal 28 volts. Protection is provided against a continuous bus voltage up to 50 volts by an internal limiting circuit which is activated at about 29 volts. The limiter is interconnected with the 12 volt regulator which supplies power to all stages but the power amplifier. The circuit board and schematic are shown in Figure 18. The picture does not show the 2N174 which is mounted elsewhere (see Figure 12) to obtain a better heat sink.⁷ This transistor, which is in series with the positive supply lead, begins to cut off when the bus voltage is excessive. It is a germanium device (for low V_{CESAT}) that has a rated dissipation of 150 watts and will therefore not



A



B

Figure 18. Limiter-Regulator Circuit

be damaged by any forward bus voltage likely to be encountered. In the reverse direction, the maximum safe voltage which may be applied is limited by V_{BEV} which is rated at 50 volts minimum.

The 12 volt Amelco regulator, is connected so that the base current of the 2N174 is used in as part of the current supplied to the low voltage bus. This reduces the load on the regulator. The power wasted in going from 28 volts to 12 volts is approximately 1.2 watts.

If the bus voltage goes below 28 volts, the transmitter power output will decrease, but no other degradation in performance will occur until the bus reaches about 21 volts, where the 12 volt regulator begins to stop regulating. Thus, the working voltage for the transmitter is 21 to 50 volts. Above 29 volts, no significant increase in total current occurs but the 2N174 will dissipate all the power above that consumed by the transmitter at 29 volts. Below 29 volts the dissipation of the 2N174 is about .2 watts and the efficiency of the entire regulator-limiter circuit is 95%.

Line Filter

An RFI line filter is built into the transmitter. It consists of an Erie filter-con for each power lead and a low pass Tee network composed of two ferrite bead inductors and 100 Pf capacitor for each data lead. Care had to be taken in filtering the data leads to prevent attenuation of the modulating frequencies. The filter components are mounted on a shallow aluminum box immediately

behind the power connector and all leads entering the box are shielded from those leaving it. The filter can be seen in the composite transmitter picture shown in Figure 24.

RFI/EMC TESTS

Extensive RFI/EMC tests were performed on the qualification transmitter by the Bendix Field Engineering Corporation, Owings Mill, Maryland in the frequency range from audio to 1.2 GHz. Detailed information regarding these tests is available in their report⁸, the principal findings of which are summarized below.

All wideband radiation and susceptibility measurements were in spec. and except at several discrete frequencies, which are discussed below, all narrow-band measurements were in spec.

A 5 MHz signal (from the VCO) on the order of 55 db above 1 microvolt (55 db uv) is conducted from the transmitter on the power leads resulting in a worst case specification violation of 21 db. The radiated level of this signal is in spec. (less than 21 db uv). Tests for susceptibility in the audio and 5 MHz regions indicated some susceptibility but not enough to cause any specification violation. Tests for susceptibility outside these two regions were in spec. by a wide margin.

The transmitter did not meet the rather stringent specification regarding radiated energy at the fundamental and harmonic frequencies. The worst case

violation, as might be expected, occurred at the transmitter fundamental frequency. The specification allows a field strength of 34 db uv maximum outside the transmitter case. At first, the measured value was 77 db uv, 43 db out of spec. Then, shielding around the output connector was improved by adding aluminum shim material and a flexible conductive pad inside the transmitter between the chassis wall and the cover. Also, an aluminum plate with a conductive pad underneath was attached outside the cover, concentric to the output connector. This reduced the radiated field to 60 db uv, 26 db out of spec. Additional experimental shielding with metal foil did not help.

It was then decided to measure the leakage from a 6 foot piece of solid aluminum coaxial cable in order to establish a frame of reference from which to judge the transmitter. With 10 watts fed from a source outside the shielded test room through the cable into a dummy load also outside of the test room, the field strength near the cable was 29 db uv, 5 db within spec. The same test performed on the OAO spacecraft cable, RG9AU (double shielded) on which the connectors had been carefully installed indicated a field strength of 54 db uv, 20 db out of spec! Thus, although the transmitter is out of spec., the field strength contribution in the spacecraft from it is not much greater than that due to the cable. RFI systems tests which have subsequently been performed with the transmitter on the spacecraft have indicated no RFI problems caused by the transmitter.

CONCLUSION

Three 10 watt FM transmitters have been designed and developed for the OAO-B Satellite Project. The performance achieved in comparison to the design specifications is shown below. The figures referred to in parenthesis by some of the specifications depict a plot of that specification versus temperature.

OAO WIDEBAND TRANSMITTER SPECIFICATIONS

Temperature Range: 0-130°F (-18°C to +55°C)

<u>Specification</u>	<u>Measured (Typical)</u>
Power Output: 7 watts min.	10 w, 8 watts min. (Figure 19)
DC to RF Efficiency: 30% min.	35% average (Figure 19)
Load VSWR Permitted: 2:1 max.	Infinite
Frequency Stability: $\pm 0.003\%$	$\pm 0.0015\%$ (Figure 20)
Non-Harmonic Spurious Outputs: -80 db	-60 db
Second Harmonic Output: -60 db	Greater than -75 db with filter
Analog Deviation: 125 KHz ± 10 KHz for 4 v PP input	125 KHz ± 5 KHz for 4 v PP input (Figure 21)
Analog Frequency Response (3 db): 20 Hz max. to between 66 and 75 KHz	2 Hz to 68 KHz
Total Harmonic Distortion: 3% max.	.5% (Figure 21)
Incidental FM: 1% max.	.3%

<u>Specification</u>	<u>Measured (Typical)</u>
Incidental AM: 5% max.	2.5%
Digital Deviation: -100 KHz; -10 KHz, +0 KHz tolerance	-100 KHz; -5 KHz, +0 KHz tolerance (Figure 21)
Digital Frequency Response: DC to between 66 and 75 KHz	DC to 72 KHz
Digital Triggering Voltage: 3.5 to 5 volts	4.3 volts
B+ Telemetry Voltage: 7.8 \pm 2 volts	7.8 volts (for 28 v bus)
Modulation Present Voltage: 8 volts \pm 2 volts	8.4 volts
RF Telemetry Voltage: 2 to 5 volts	3.6 v for 10 watts output
Transient Voltage Allowed: \pm 20% of 28 volts	+78, -390%
Weight: 3.9 Pounds max.	3.3 pounds
Dimensions: 8-3/4 \times 4 \times 2 inches	8-3/4 \times 4 \times 2 inches

Only one of the above specifications was not met. Nonharmonic spurious outputs were 60 rather than 80 db below the unmodulated carrier. All other specifications were met or exceeded. The frequency stability achieved, \pm 0015%, exceeds the specification by a factor of two and is particularly significant in view of the difficulty of achieving this performance in a wideband FM transmitter.

Other salient features embodied in the transmitter include

1. Will operate continuously into an RF open or short circuit without damage

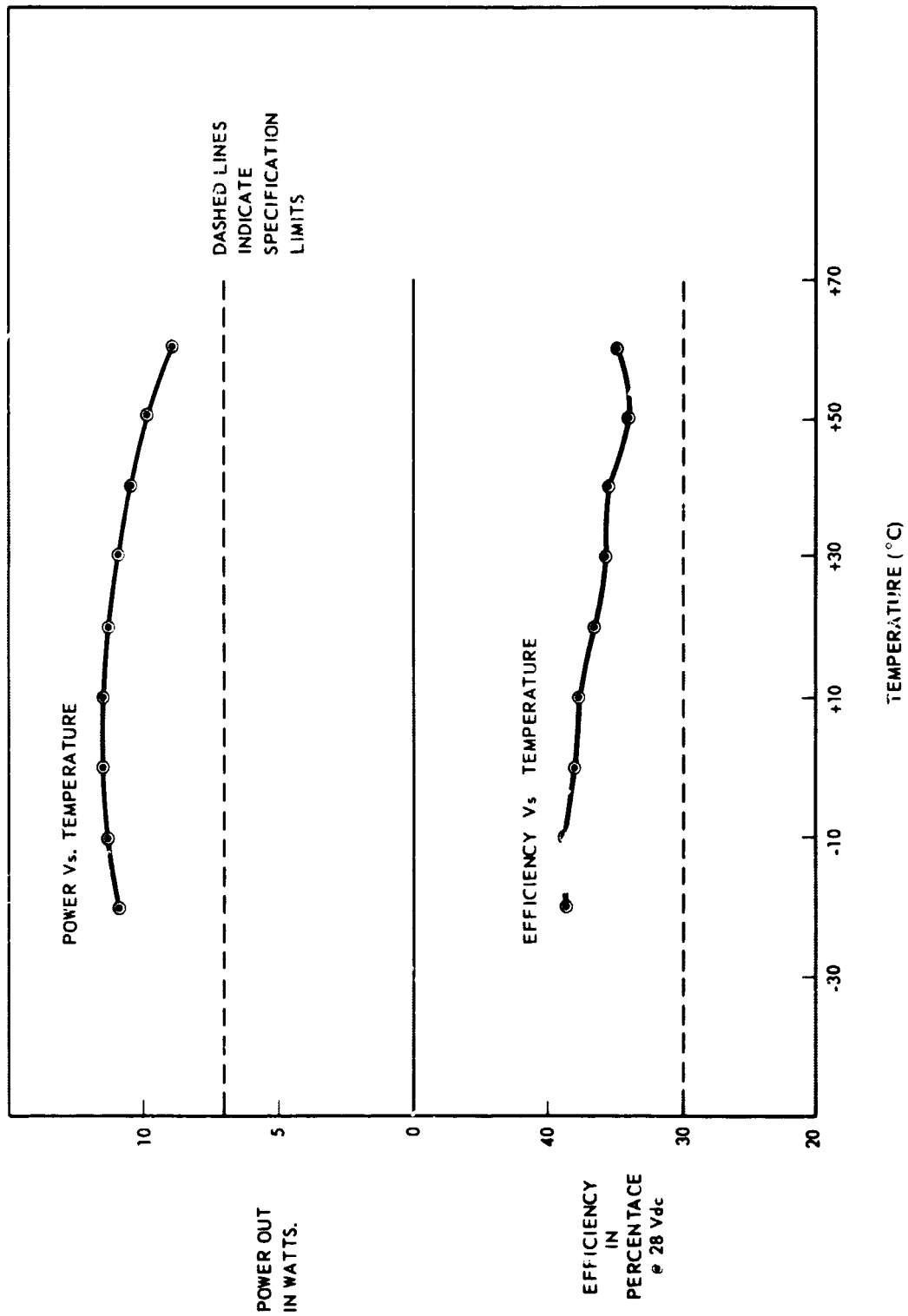


Figure 19

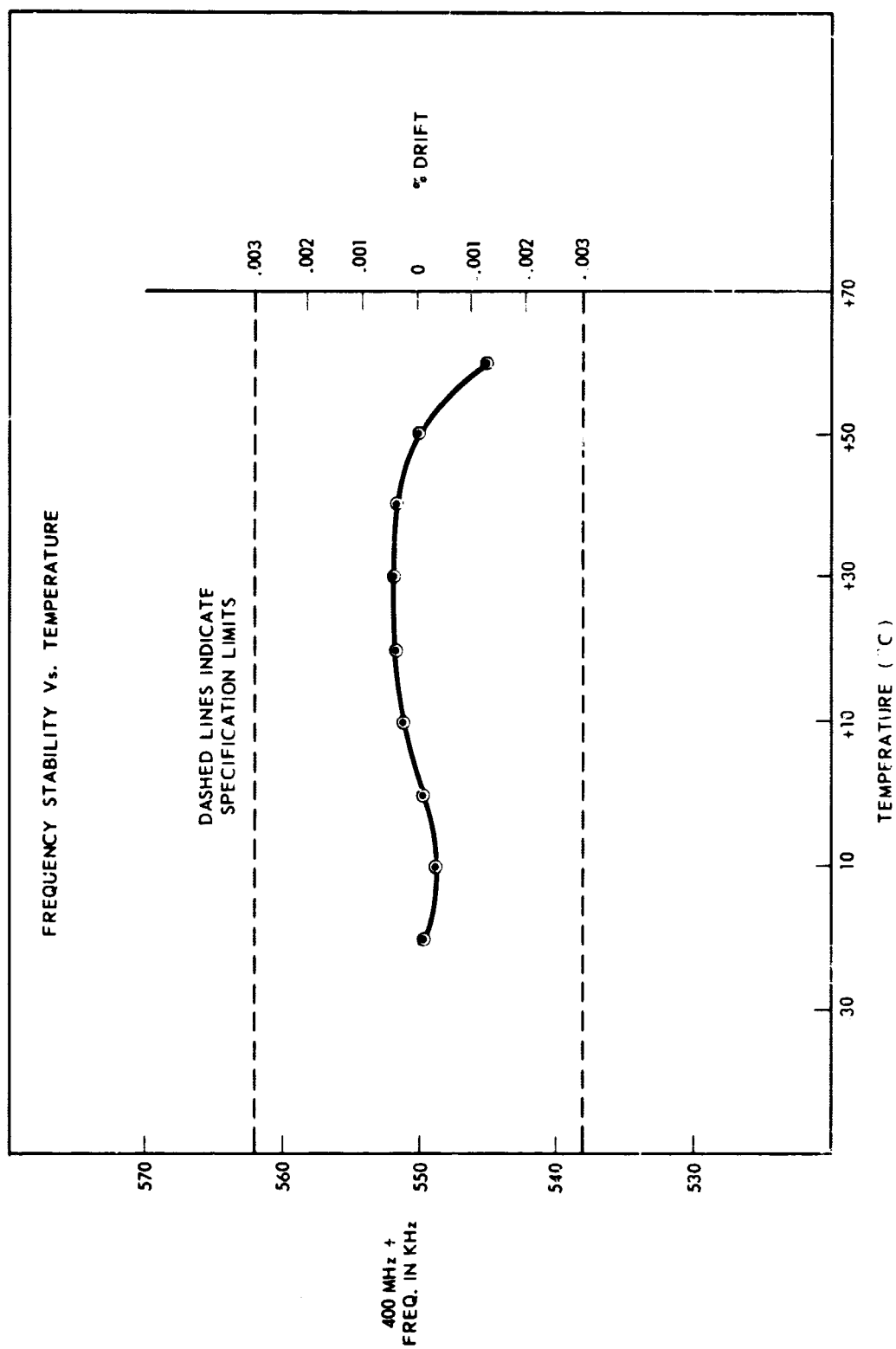


Figure 20. Frequency Stability Versus Temperature

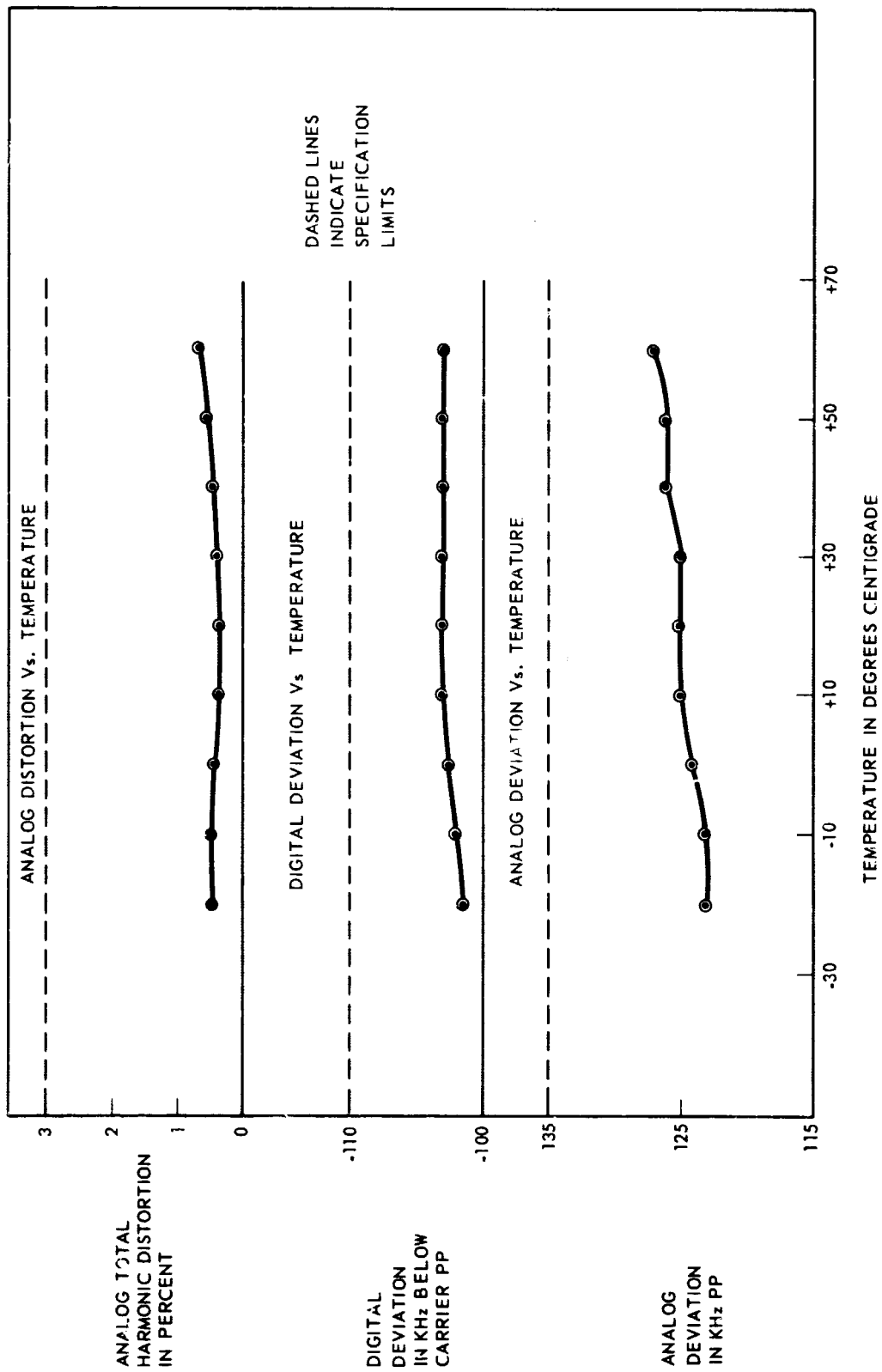


Figure 21

2. Purely resistive 50 ohm output impedance
3. Package efficiency 35% with load isolator
4. Deviation capability of ± 200 KHz at 1% total harmonic distortion
5. Modulating frequency range from DC to 175 KHz
6. Protection against positive and negative transients and sustained supply voltage up to 50 volts
7. Does not require a DC to DC power converter.

A block diagram and a complete schematic of the transmitter are shown in Figures 22 and 23. Photographs are shown in Figures 24, 25 and 26.

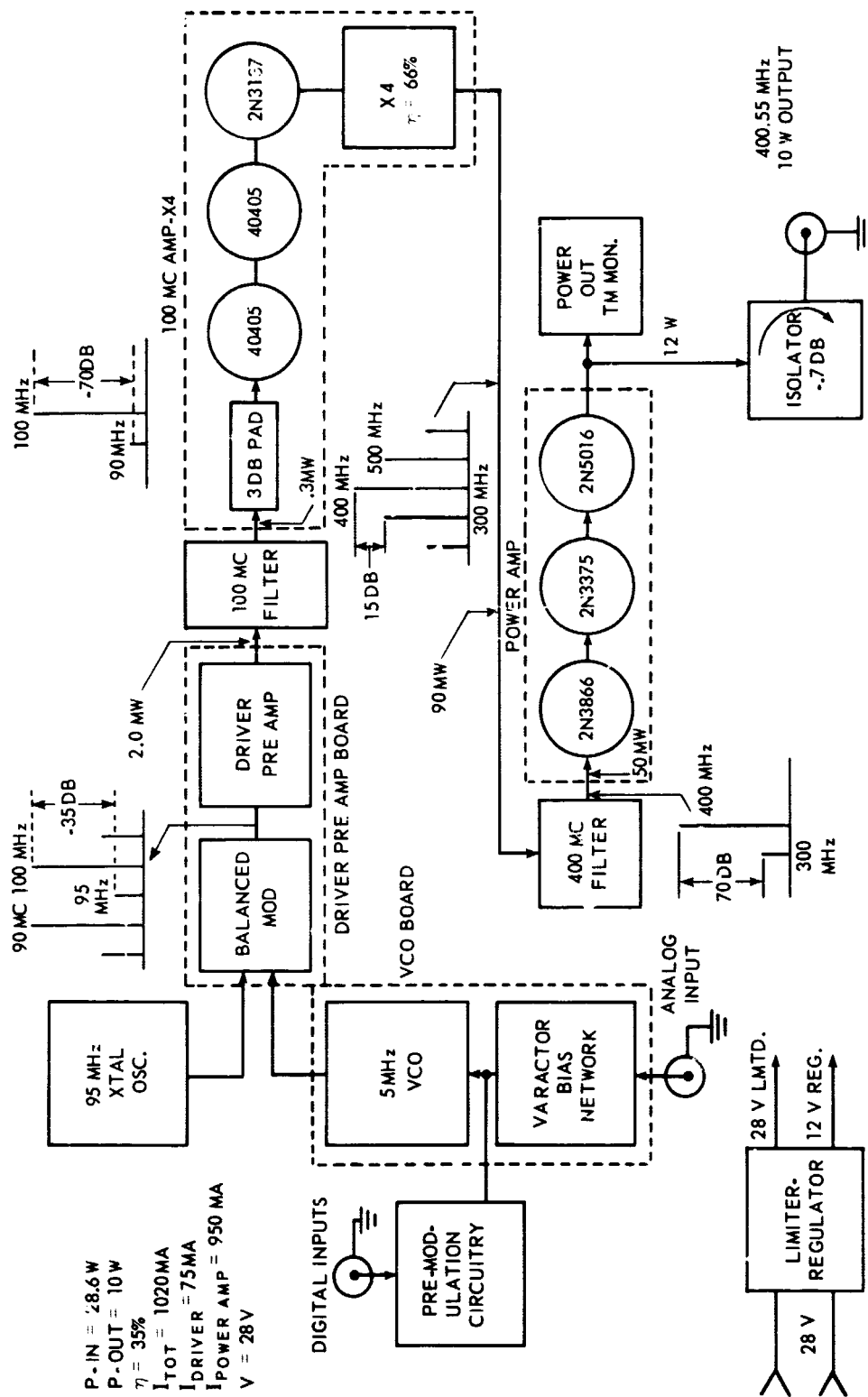


Figure 22. Block Diagram OAO Wideband Transmitter

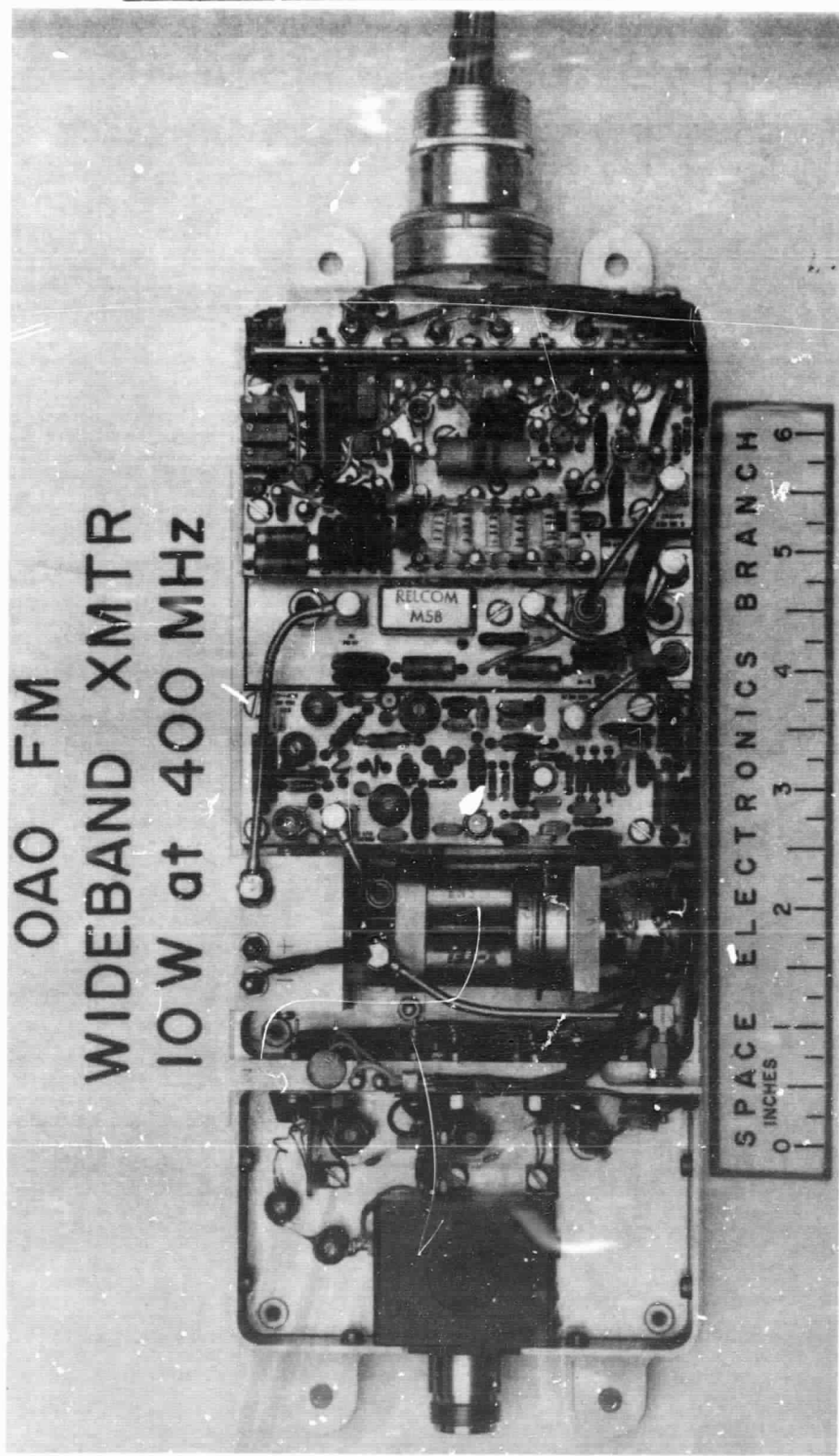
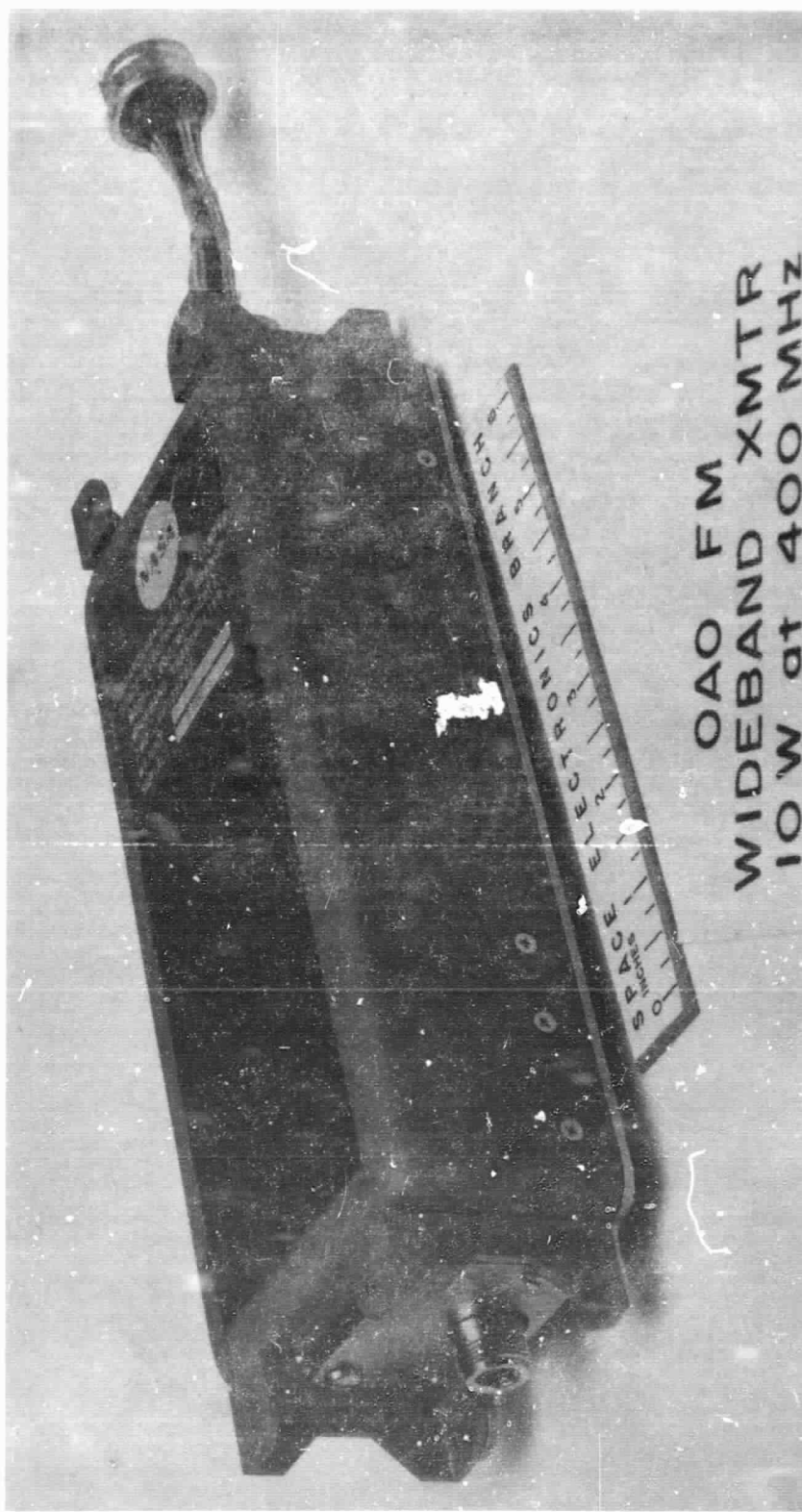


Figure 24



OAO FM
WIDEBAND XMTR
10 W at 400 MHz

Figure 25

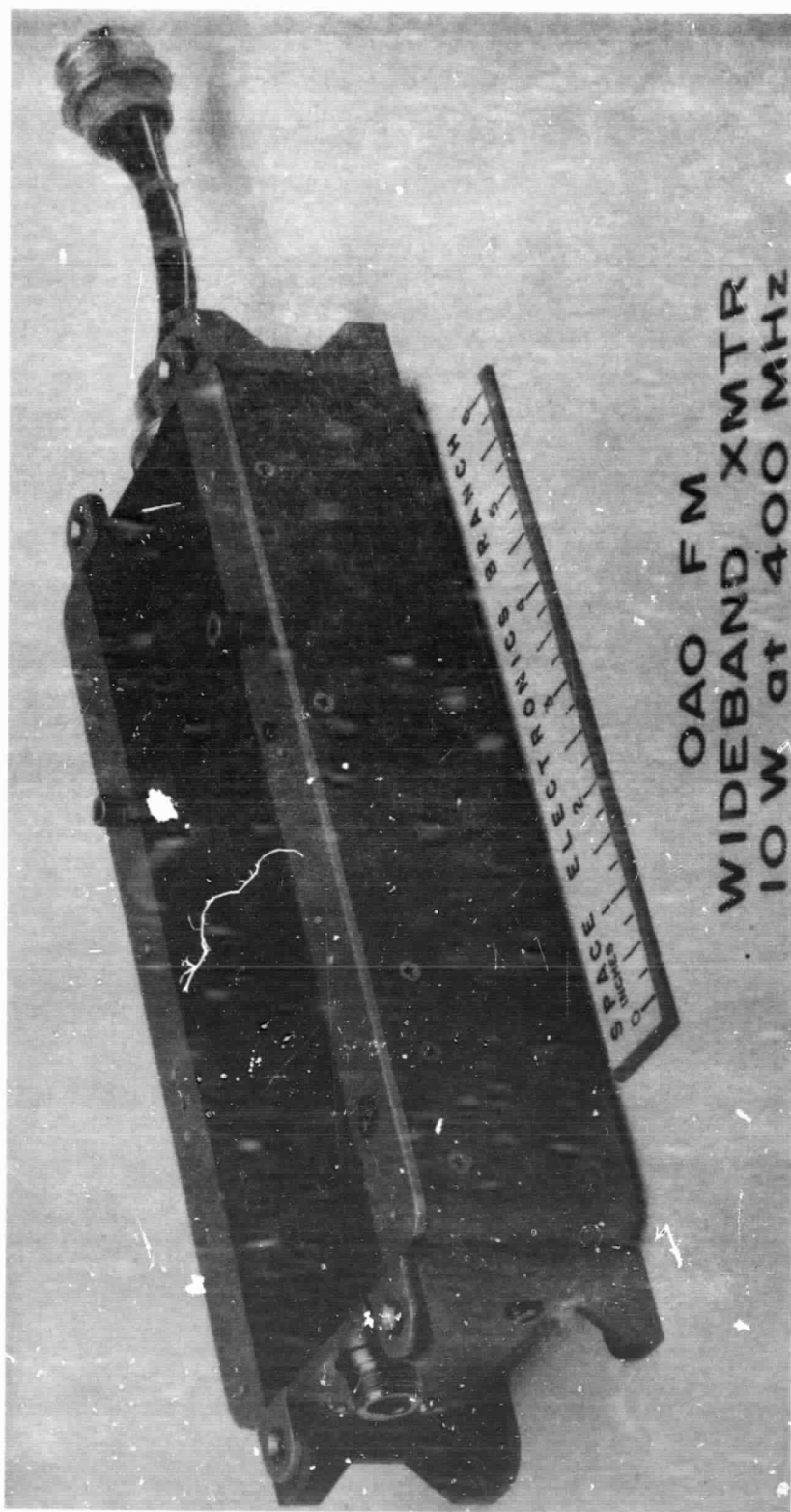


Figure 26

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APPENDIX

The recent development of hyperabrupt varactors which have a capacitance ratio $\left(\frac{C_{2\text{VOLTS}}}{C_{10\text{V}}}\right)$ greater than 10 and a characteristic voltage versus capacity curve that varies inversely with the square instead of the square root (which is more common) of the applied voltage has made it possible to obtain linearity in a VCO with an extremely simple circuit. If one of these varactors such as a Motorola MV1403 is connected in series with a fixed capacitor of the correct value, the combination capacitance will vary linearly with varactor voltage and produce a linear change in frequency over a deviation range of several percent.

That the series combination capacity varies linearly with varactor voltage is illustrated in Figure A-1. The lower curve is a plot of varactor capacity, C_v , versus bias voltage, V_B , for an MV1403. The upper curve is a plot of $C_{\text{TOT}} = \frac{C_f C_v}{C_f + C_v}$ versus C_v where C_f is a fixed capacitor of 156 pf (selected empirically), C_v is the varactor capacity, and C_{TOT} is the total capacity of C_f and C_v in series. The straight line is a plot of C_{TOT} versus V_B and illustrates the point that the series combination capacity of C_v and C_f varies linearly with V_B . This useful result occurs because the non-linearity of the upper and lower curves cancel. For example, assume we wish to find C_{TOT} when V_B equals 5 volts. First, use the lower C_v versus V_B curve to determine that when V_B is 5 volts, C_v (on the ordinate) is 75 pf. Then find C_{TOT} when C_v equals 75 pf by locating $C_v = 75$ pf on the abscissa and projecting to the upper C_{TOT} versus C_v curve and read $C_{\text{TOT}} = 50$ pf. Thus,

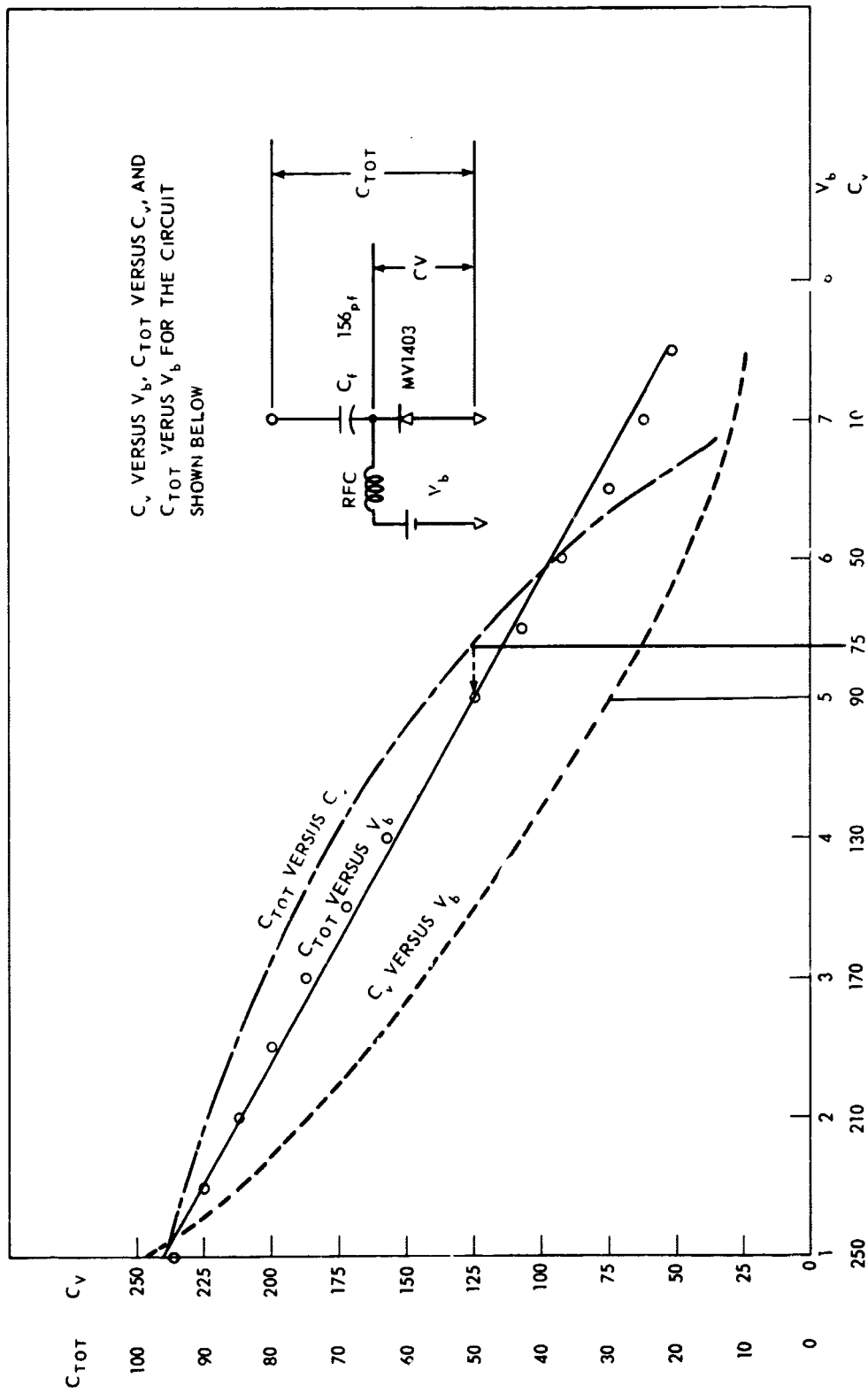


Figure A-1

when V_B equals 5 volts, C_{TOT} equals 50 pf. Note that this point falls on the straight line. Similarly, all other values of V_B and C_{TOT} will fall on or close to the straight line. As V_B varies from 1 to 7.5 volts C_{TOT} varies approximately 80 pf with reasonable linearity over the entire voltage range and with extreme linearity over a two or three volt selected region.

That a linear capacity change can produce a several percent linear frequency change is illustrated in Figures 2 and 3. The frequency of an oscillator will vary according to $f_o = \frac{1}{2\pi\sqrt{LC}}$ where f_o is the frequency of oscillation, L is the inductance and C the capacitance of the resonant circuit. If we choose $L = 1.1$ microhenries and vary C from 200 to 2000 pf, the frequency of oscillation will vary from about 3.5 to 15 MHz as shown in Figure A-2. The curve is obviously not linear over the entire frequency range. But let us assume we are only interested in varying the frequency of a 5 MHz VCO by several percent. The total value of C needed to produce oscillations at 5 MHz is approximately 920 pf. To change the frequency by 4 percent to 4.80 MHz, the capacity must be increased by approximately 80 pf. Thus, the only region of the curve in Figure A-2 of interest is that marked between 4.8 and 5 MHz. This section is expanded in Figure A-3. The horizontal axis is capacity added to 920 pf and the vertical axis is the resulting decrease in frequency. The linearity over the 200 KHz region is approximately ± 1 percent.

Looking again at Figure A-1, we see that we can change the capacity of C_{TOT} by a maximum of 80 pf. To prevent forward conduction in the varactor due to

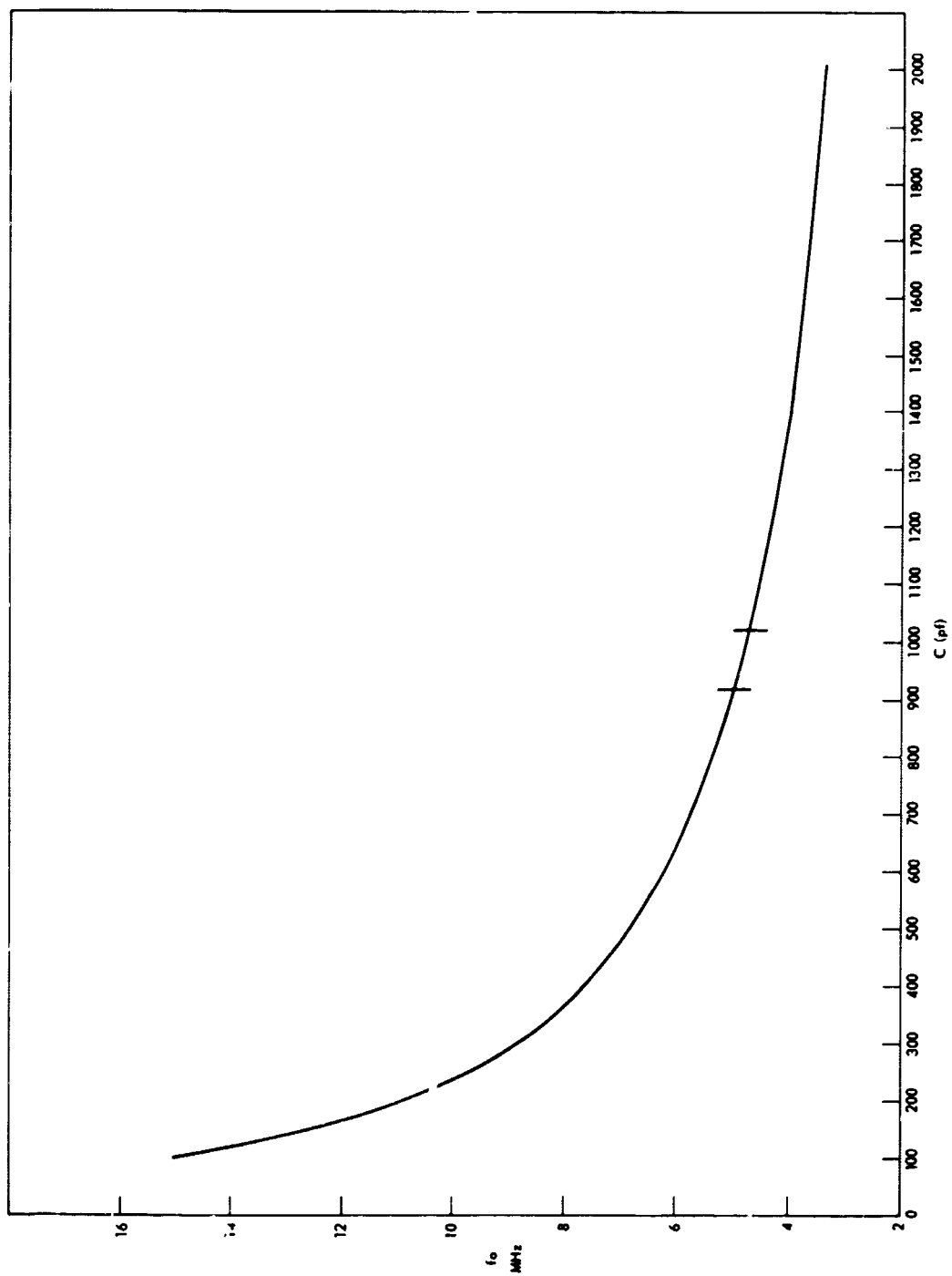


Figure A-2. Resonant Frequency Versus Capacity For an LC Circuit Having a Fixed L of $1.1 \mu\text{H}$

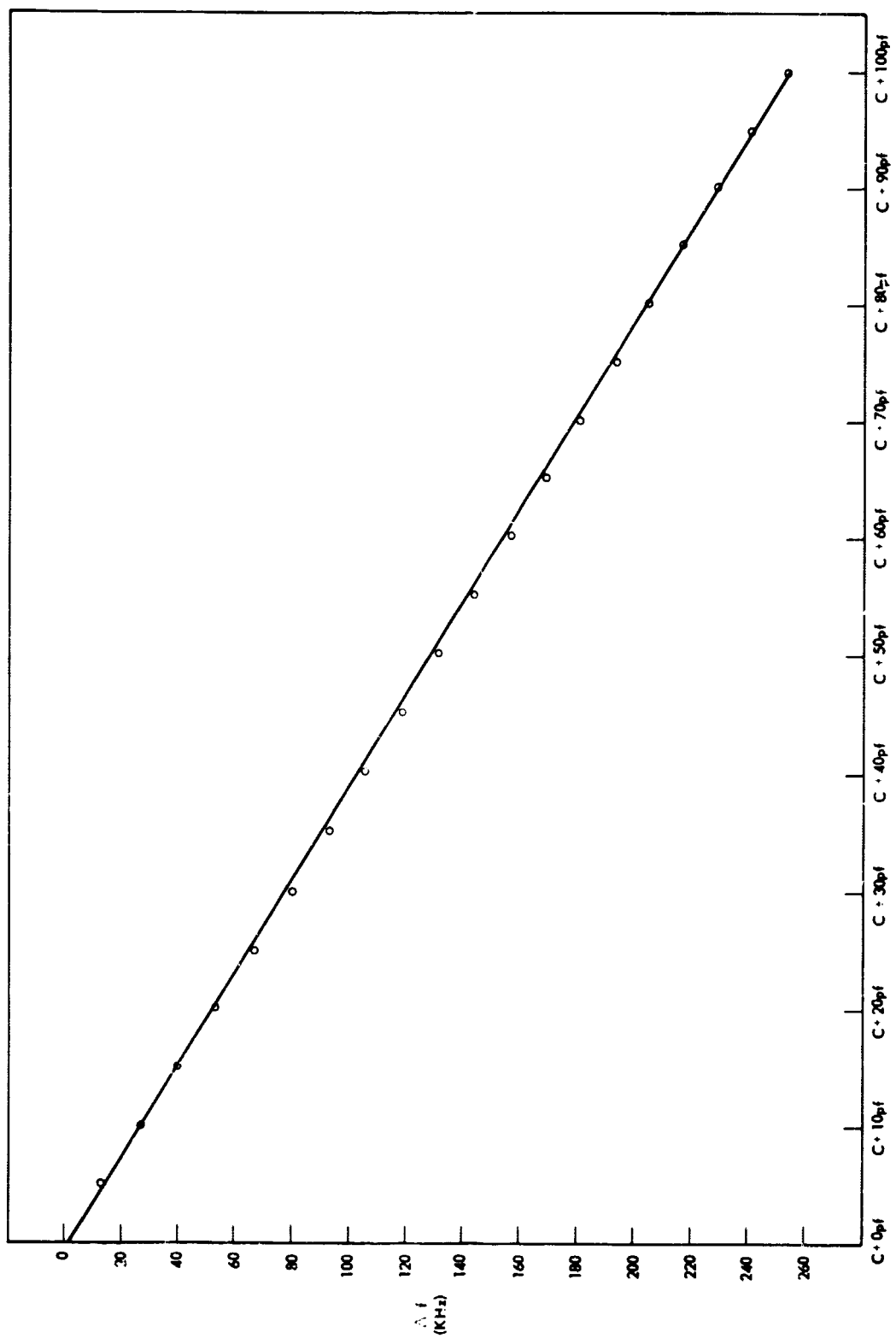


Figure A-3. Decrease in Resonant Frequency Versus Added Capacity For an LC Circuit Having a Fixed L of 1.1 μ h and a Fixed C of 920 Pf

rectified RF voltage we shall limit the lower bias voltage to 3 volts and to assure good linearity we shall limit the upper bias voltage to 6 volts. Thus, we have an operating region from 3 to 6 volts, and a corresponding linearly varying capacity of roughly 40 pf. The slope of the curve in Figure A-3 indicates that the frequency of oscillation will change 2.5 KHz per pf. Thus, the linear deviation region of the VCO will be $\frac{2.5 \text{ KHz}}{\text{pf}} \times 40 \text{ pf} = 100 \text{ KHz}$ or 2 percent. Obviously greater deviation can be obtained if one is willing to accept poorer linearity and vice versa. The most linear operating region varies from one diode to another and must usually be determined by experimental adjustment of bias voltage for minimum distortion. The total harmonic distortion typical of a 5 MHz VCO linearized by this method is shown in Figure A-4. Linear deviation greater than that discussed is possible at the cost of stability by operating on a steeper section of the curve in Figure A-2 and adjusting C_f and V_B to compensate for the curvature.

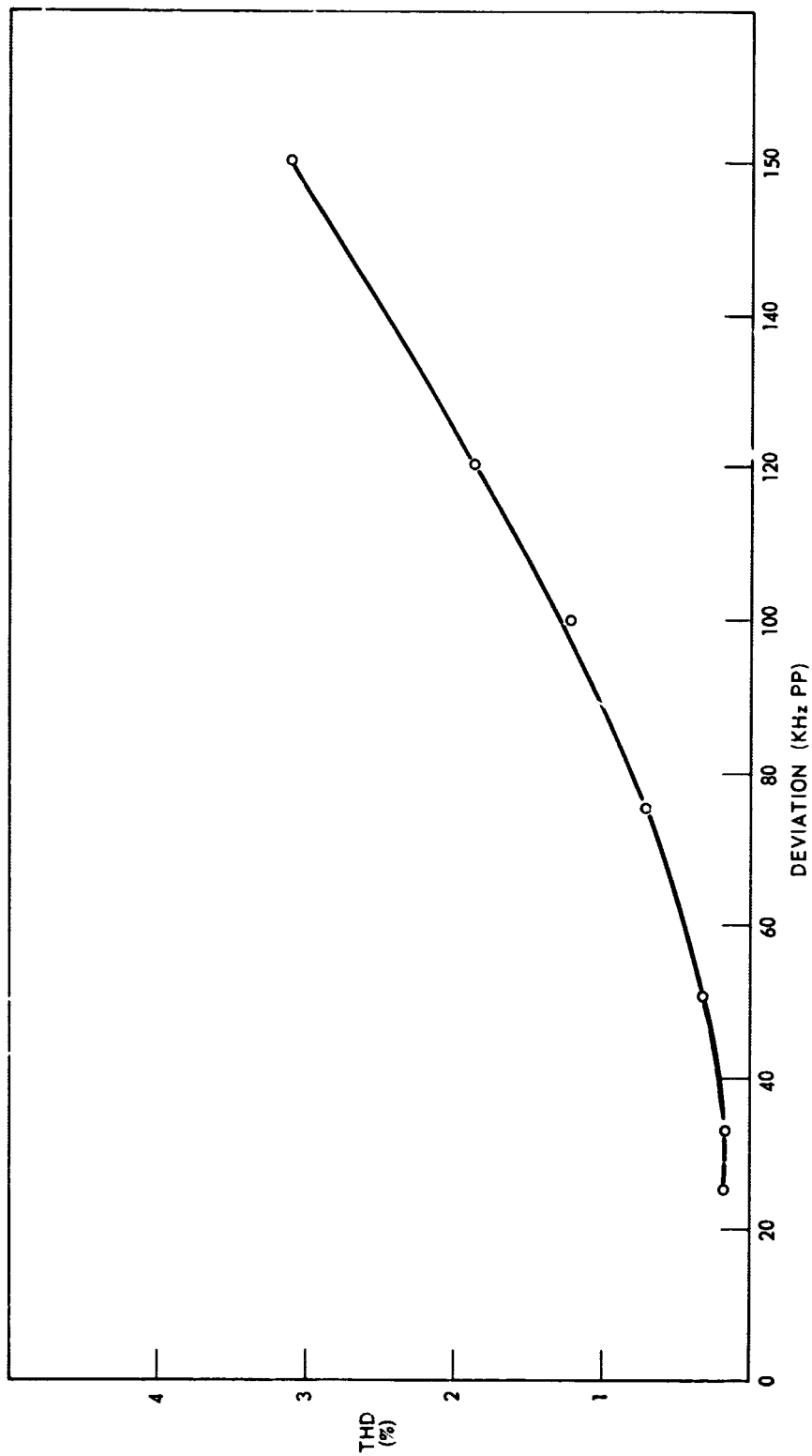


Figure A-4. Total Harmonic Distortion Versus PP Deviation For 5 MHz VCO